WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

OCTOBER 1954

VOL. 31 No. 10 THREE SHILLINGS AND SIXPENCE

Model 8

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20,000 ohms per volt *plus* AUTOMATIC Overload Protection

Produced in response to a demand for a high sensitivity version of the world-famous Universal AvoMeter, this model incorporates the traditional design features of its predecessors, so highly valued for simplicity of operation and compact portability.

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2·5V.	50μA.	2·5V.	100mA.	*First indication* 0·5 Ω.
10V.	250μA.	10V.	1A.	*Maximum indication* 20M Ω.
25V.	1mA.	25V.	2·5A.	0—2,000 Ω (using internal batteries)
100V.	10mA.	100V.	10A.	0—200,000 Ω (using internal batteries)
250V.	100mA.	250V.	—	0—20M Ω (using internal batteries)
1,000V.	1A.	1,000V.	—	(using internal batteries)
2,500V.	10A.	2,500V.	—	0—200M Ω (using external batteries)

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A8/1

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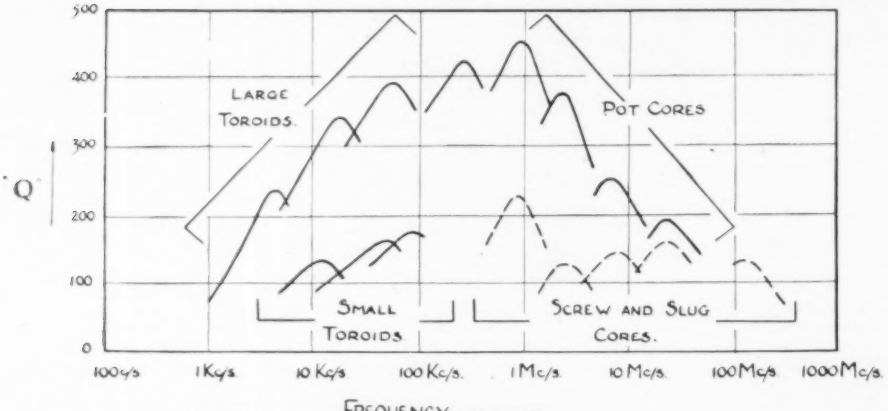

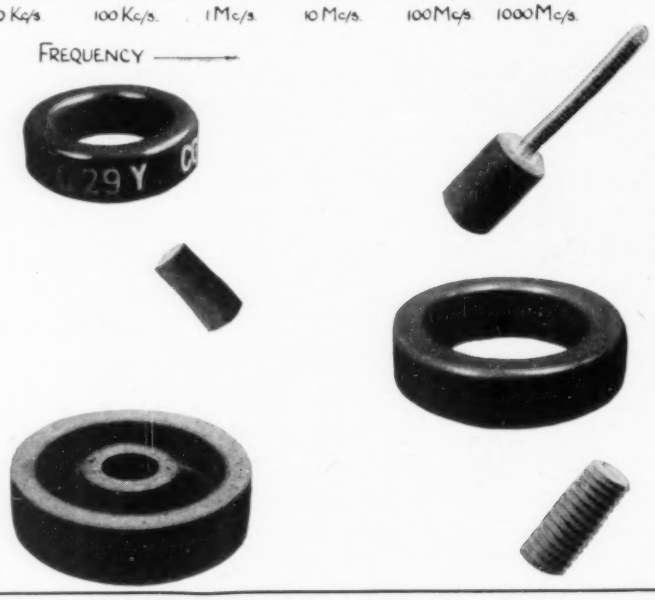

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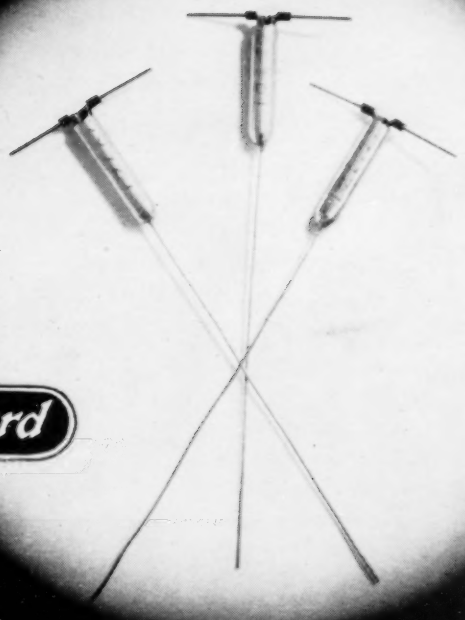

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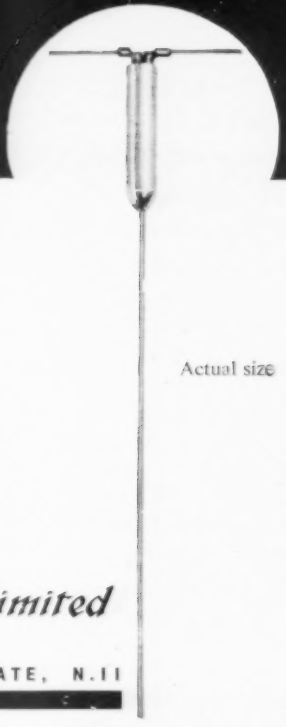

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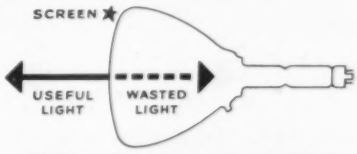

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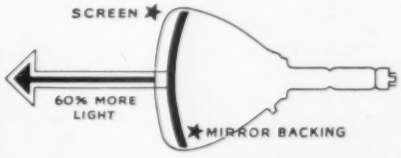
Without aluminizing, tubes waste half their light (see diagram above). To counteract this the brilliance must be increased and the tube life is shortened.



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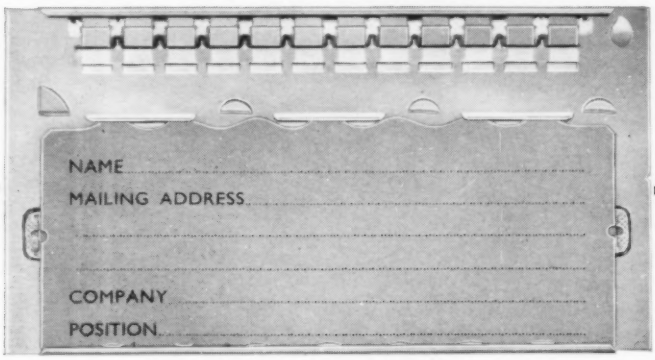

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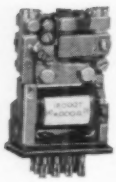

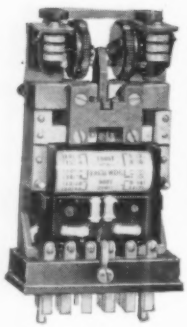

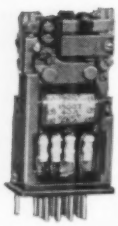

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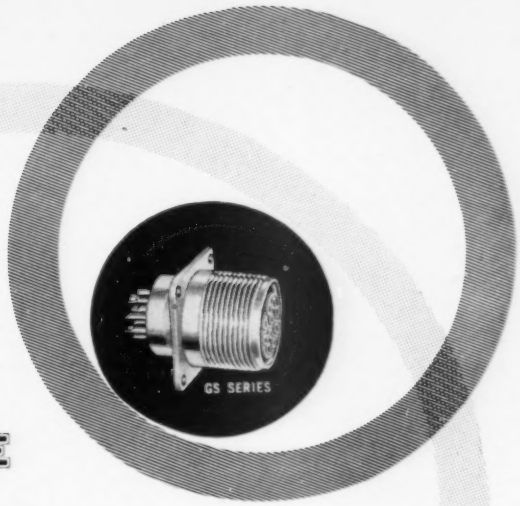

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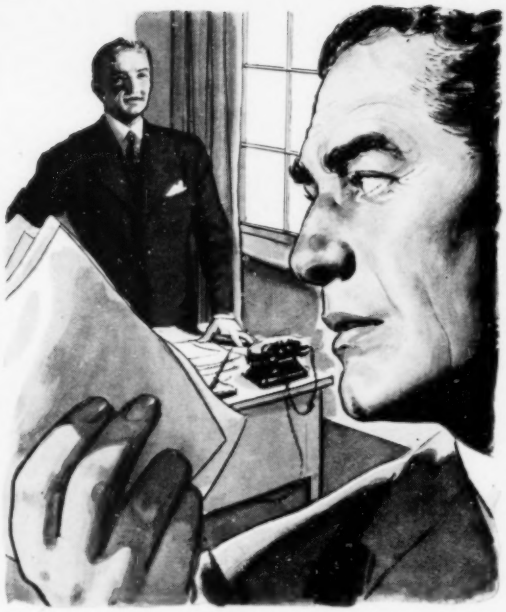

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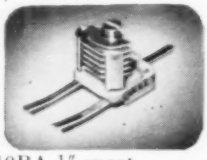

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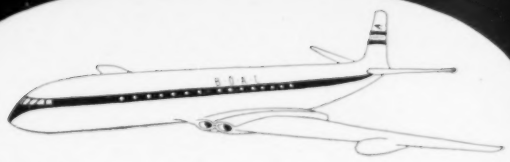

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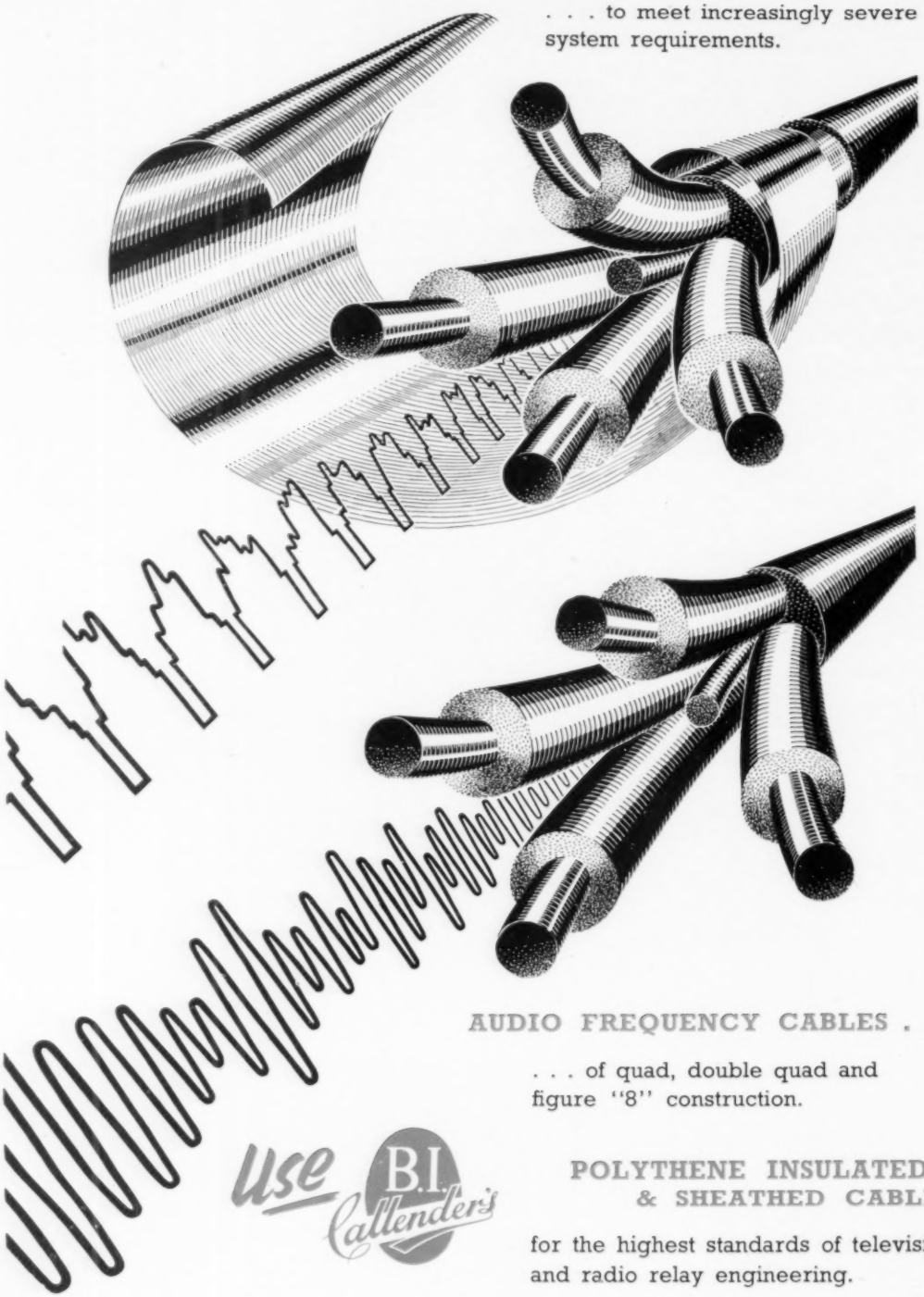

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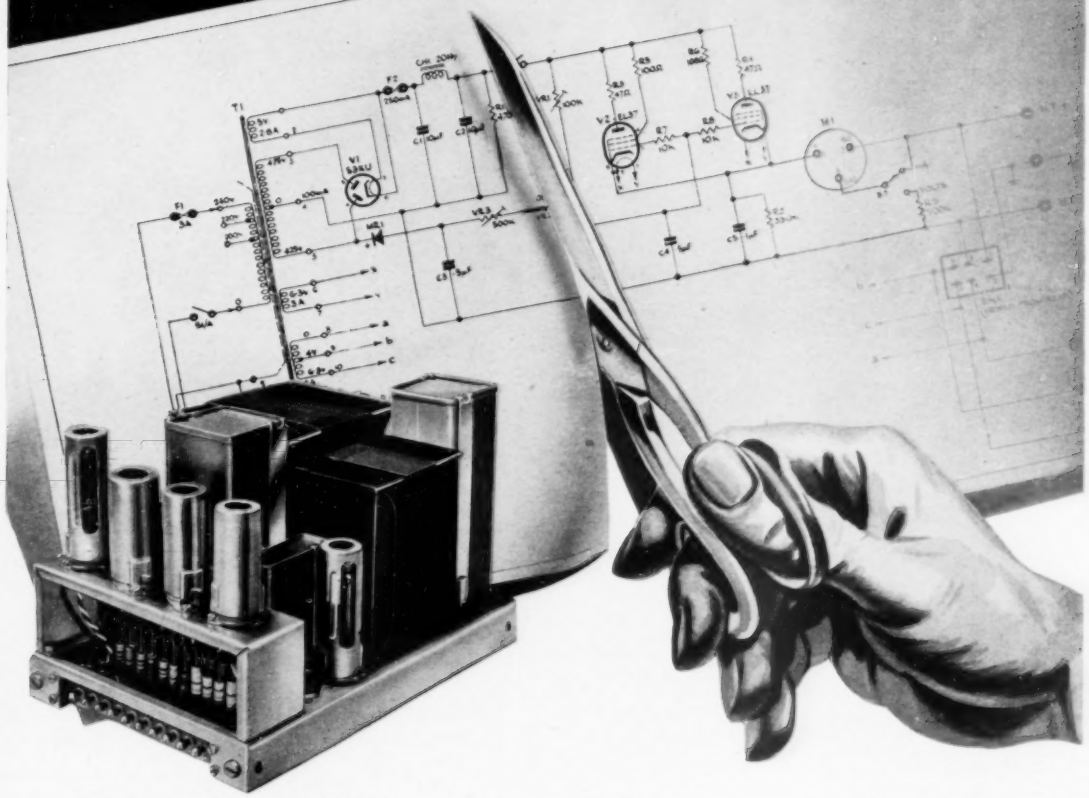

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QV06-20

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V_h	6.3 V
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V_{g1} max.	—150	V
f max.	175	Mc/s

Base Octal.

Typical Applications	V_a (V)	Pout (W)	f (Mc/s)
R.F. Power Amplifier			
Class "C" Telegraphy and F.M. Telephony	{600 {320	52 25	60 175
Class "C" Anode and Screen-grid Modulated	400	32	60
A.F. Power Amplifier or Modulator (Two valves.)			
Class "AB1"	600	82	A.F.
Class "AB2"	600	90	A.F.

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The Teaching of Electromagnetism

IN the June Editorial we discussed some of the points raised in the paper read before the I.E.E. by Mr. P. Hammond on electromagnetic theory. The paper has now been published in the July number of the *Proceedings* (Part I) and with it a number of communications to the discussion. These should certainly be carefully studied by all interested in the teaching of any branch of the subject, if only to get some idea of the diversity of outlook and opinion—not to say muddle—that pervades the fundamentals of the subject, and the teaching of these fundamentals to students. To illustrate this we will give a number of quotations from the various contributions:—

1. "One should not accord to the inverse-square law a dignity greater than that of a stepping stone. It is hard to find a definition of the word 'law' that would justify its application to the inverse-square relation . . . So the inverse-square law is a statement about geometry, and not about electricity, magnetism or gravitation. To pretend otherwise is to mislead students."

2. "A law is something fundamental, while a rule is not. The theory of electrical machines is normally based on the flux-cutting rule, but doubtful points can be settled by going back to the more fundamental flux-threading law . . . . The Faraday disc is well known as an apparently simple apparatus which requires some very difficult theory to explain it. It is a topic more suitable for correspondence in the electrical Press than for an undergraduate student's course, and both it and the anomalies to which it leads should be omitted from the syllabus." The same writer very rightly says that "what we call a 'rotating flux' neither flows nor rotates."

3. "Is the author so convinced of the virtues of symmetry that he is wise to build his system on another concept which is basically irrelevant to the scheme of things—the isolated magnetic pole?"

4. "However deeply we may respect Maxwell and Faraday, we must admit that they have had a bad effect on our attitude towards electrostatics and magnetism, since so much of their work is coloured by the desire to explain phenomena in terms of stresses in the ether . . . . I would draw attention to a defect which the author's treatment has in common with the teaching methods put forward in connection with M.K.S. units, namely an entire neglect of the role of magnetic polarization of iron . . . . In some difficult cases of electromagnetic induction (e.g., Faraday's disc) I find it helpful to regard the e.m.f. as the integral of the forces Bev on the electronic charges e moving with velocity v relative to the system producing the magnetic field." This is a point that we stressed in the June Editorial.

5. "All practising engineers respect theory, but in a detached way, much as they might have regard for astronomy; this is because their direct concern is not with theory but with their possible use of it, and the paper does no service to such engineers if it asks them to respect this or that theoretical interpretation for no more reason than that of removing asymmetries from some equations they never use . . . . The electric-charge approach I have sketched is classical . . . it puts magnetism in a proper perspective—that of a very convenient but unessential concept."

6. "The author suggests that the student should not be taught to ask whether the flux

linkages are changing. I must disagree strongly with this. The fundamental is a flux linkage, not a flux as such, nor a linkage as such, and induced e.m.f. is a dual function of that fundamental."

7. "A separate issue is the teaching of the history of electromagnetism, its units and dimensions, and I think this is as important as teaching any particular theory. If history is taught, the student will learn that the subject of electromagnetism is not closed, and he will appreciate that there is still much in the fundamentals for research."

8. With reference to the Faraday disc without any external circuit, one contributor says, "If the charge segregation is produced by relative motion of the magnet and disc, it seems immaterial which is actually rotated. On this point Einstein is explicit: 'The observable phenomenon . . . depends only on the relative motion of the conductor and the magnet, whereas the customary view draws a sharp distinction between the two cases in which either the one or the other of these bodies is in motion' . . . . This, then, suggests that the field of a magnet does rotate with it, inasmuch as the rotation would be expected to produce effects in stationary bodies, even when the field is uniform."

9. "I prefer to regard all magnetic effects as due to electric currents, and I think that his excuse for introducing the equation $F = m_1 m_2/(\alpha r^2)$—that it brings mathematical symmetry into the system—is an inadequate reason for distorting the physical system . . . . There seems to be no reason why μ_0 and κ_0 could not be dispensed with and appropriate use made of c only. μ_0 is $120\pi/c$ and κ_0 is $(120\pi c)^{-1}$."

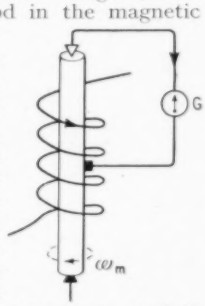
10. "My view is that the practical importance of permanent magnets nowadays fully justifies the inclusion of magnetostatics in an electrical engineering course . . . . It is pleasing to note that some teachers now recognize the drawbacks of distorting the presentation of basic theory to accommodate a particular unit system."

11. "Although I am a keen exponent of the M.K.S. system, I am always accustomed to relate it, first and foremost, to the inverse-square laws . . . . The opinion that, in the M.K.S. system, the inverse-square law 'tends to become a mere dustbin for unwanted constants' is one against which I protested when it was originally put forward."

12. "Relativistic electromagnetic theory provides a definite answer to the old puzzle of whether the e.m.f. is induced in the rotating magnet or in the stationary wires . . . . It may be shown that when a permanent magnet moves with rectilinear velocity v it causes an electric field, without the displacement of any free charges in it, given by $\mathcal{E} = \mu_0(H \times v) + J \times v$, where H has the magnetostatic meaning and J is the intensity of magnetization. This can be applied to a rotating cylindrical magnet if H is negligible, i.e., if the magnet is long, for the term $J \times v$ depends only on conditions in an element of the magnet coincident with the stationary point where the electric field is $\mathcal{E}$, whereas H depends on all elements and their individual linear velocities. For a long permanent magnet $J \simeq B$ so that $\mathcal{E} \simeq B \times v$ and the force on a unit charge, in and rotating with the magnet, is $F = \mathcal{E} + v \times B \simeq 0$. Thus no e.m.f. is induced in the magnet, so that the e.m.f. must be induced in the stationary leads to the galvanometer."

We do not pretend to understand the reasoning of this last quotation, but the result is such as to convince us that there is something wrong with it. It may be the application of relativistic formulae based on a rectilinear velocity v to a body which has no such velocity, but is rotating with an angular velocity ω. As we have previously emphasized, v is relative to some frame of reference, but ω is absolute. Leaving all mathematics aside and looking at the experiment from a commonsense point of view, we have a wire circuit and galvanometer at rest in a magnetic field which is undergoing no change whatever either in strength or direction, and we have what is virtually a generator, consisting of a cylindrical conductor being rotated in a magnetic field. It seems almost incredible that anyone can seriously maintain that the e.m.f. is induced in the external circuit and not in the primitive generator, but two or three contributors to the discussion appear to support this view, and in every case based on what we consider to be misapplications of relativistic theory.

If, instead of the permanent magnet, one rotated a circular copper rod in the magnetic field of a coil, as shown in the figure, current would flow through the external circuit as shown, and surely no one would suggest that the e.m.f. is induced in the external circuit. In the case of the permanent magnet the magnetic field is very similar and equally stationary and unchanging. If we picture one of the free electrons in the rotating copper or steel, can we imagine it possible for it to know whether the magnetic field in which it is moving is the result of an external solenoid or of the myriads of molecules which constitute the magnet, and in the latter case to refuse to obey the laws of electromagnetism and experience a force? It is equally fantastic to expect the electrons in the external circuit to discriminate between the fields produced by the solenoid and the permanent magnet.



If the e.m.f. is $E\omega_m$ when the angular velocity of

the magnet is ω_m and the external circuit at rest, then it will be $-E\omega_m$ if the magnet is at rest and the external circuit rotating in the same direction at the same angular velocity. If ω_c of the circuit is equal and opposite to ω_m the e.m.f. will be $2E\omega_m$, and in general, e.m.f. $= E(\omega_m - \omega_c)$. Although one can refer to this difference between the two angular velocities as a relative angular velocity, one must not confuse it with 'relativity' as applied to linear velocities.

If there is no external circuit, charges will distribute themselves on and in the rotating magnet in such a way as to neutralize the induced electric force at every point within the magnet. These charges will produce an external electric field, and it is this that produces the effects in stationary bodies referred to by one contributor, and not the rotation of the magnetic field.

We hope that this simple non-mathematical treatment of the question will go some way towards removing the clouds which have been cast on it by the shades of Einstein.

G. W. O. H.

L.F. NOISE IN RESISTORS

Excited by R.F. Current

By Noel B. Montagnon

SUMMARY.—In these measurements, two signals, differing in magnitude by a factor of approximately 10^7, must be separated before the noise can be observed. Several methods which were tested are described, and also some measurements which were made with the most successful methods. Noise, at 5 kc/s, from several resistors, was measured for applied frequencies between 50 kc/s and 5 Mc/s, at voltages of 40, 60 and 80 V r.m.s. This noise is found to differ in several respects from that caused by direct current. It depends inversely on a power of the applied frequency up to a few Mc/s where it becomes independent of frequency. The noise, in mean square volts per cycle bandwidth, depends on a power of the applied voltage which varies between 2·26 at 50 kc/s and 2·75 at 1 Mc/s. Some composition-type resistors exhibit an increase in noise output after an exposure to alternating current for long periods.

Introduction

THE problem of noise voltages generated spontaneously in circuit elements has been studied extensively during the past 30 or 40 years, but, with the rapidly expanding development of electronics, it is attracting more attention and has become a matter of great importance. The reason for this increased preoccupation with the subject lies in the fact that these noise voltages determine the limit of useful sensitivity of amplifiers. For signals only slightly greater than this noise level (irrespective of any outside interfering signal) at the amplifier input, an increase of gain in the amplifier does nothing to improve the intelligibility of the signal.

The generation of random voltages across circuit elements is a function of the internal structure of these elements and of the particle nature of electricity. For instance, shot noise, predicted by Schottky,[1] in 1918, is the result of the emission of electricity from the cathode in small finite amounts—like the stream of shot from an erratic machine gun rather than a continuous flow, as in a jet of water. Thermal agitation noise (Johnson[2]), which appears across the terminals of any conductor, can be shown to be produced by the random thermal motions of the free electrons in the conductor.

MS accepted by the Editor, September 1953

Two other classes of noise; flicker noise, which appears in valves having oxide-coated cathodes, and current or Bernamont noise, resulting from the flow of current in a semi-conductor, are not so easily explained, but there is no reason to believe that their cause does not lie within the material itself. There seem to be at least two reasons why these two types of noise should be regarded as arising from the same mechanism. One is that a semi-conductor is always involved and the other is that, while shot noise and thermal noise have a uniform spectral distribution, these types do not, but vary in some way with the frequency of measurement. Many investigators have studied these phenomena since their first discovery (current noise in 1919 by Kawamoto[3] and flicker noise in 1925 by Johnson[4]) and theories regarding their production and behaviour have been advanced by them and others. Tables 1 and 2 list some of these investigators with their findings and theories.

It is obvious from the above remarks that the complete elimination of noise is impossible, but the study of the subject is worth while for two widely different reasons. First, it can provide clues as to the structure of matter and to the mechanism of conduction. Secondly, it can show ways to minimize the effects of noise through the use of components and circuit designs which

reduce the noise. As an example of this, the recent work of Campbell[5] and Pavich,[6] working under Chipman at McGill University, shows that carbon resistors of different composition and construction have widely different noise-producing capabilities.

The Effect of Alternating Currents in Resistors

It seems apparent, from a study of past work on the subject, that all the experimental work and theoretical discussion has had to do with noise produced by direct current in the resistor. Any mention of the frequency dependence of the noise has referred to the frequency of measurement. At McGill University, under the auspices of the Defence Research Board of Canada, a project was under way to investigate the current–noise characteristics of commercial resistors. In the third year of this project, it was felt that an attempt to determine the effect of alternating currents would be worth while. The writer became interested in this and decided to make it the major part of his work.

At the outset, it was appreciated that such an investigation would present considerable problems of technique. Most of the available time was spent in overcoming these problems and in exploring various methods. As a consequence, the experimental results are limited in number and insufficient to form the basis of a thorough theoretical discussion. This paper deals with the methods attempted and presents the results obtained with the most successful of these.

The Approach to the Problem

In the problem of measuring noise from a resistor carrying alternating current, one of the first considerations is the order of magnitude of the impressed alternating voltage (hereafter called the exciting voltage or signal) required to

TABLE 1

EARLIER INVESTIGATIONS OF CURRENT NOISE

Investigator	Date	Materials investigated	Behaviour	Frequency
Bernamont[7–11]	1934 to 1937	Thin films of tungsten and platinum Various forms of carbon	$b = 2$ (low current) < 2 (high current) $c = 1$ (low freq.) ≈ 2 (high freq.)	96 to 162,000 c/s
Otto[12]	1934	Various forms of carbon Copper oxide rectifiers	$c = 1$ to 1·4 $d = 2$. $a = 1$	5 to 1,000 c/s
Meyer and Thiede[13]	1935	Thin carbon films	$b = 2$ $c = 1$ to 2	—
Christensen and Pearson[14]	1936	Various forms of carbon Metallic films	$a = 1$ to 1·42 $c = 1$	—
Harris, Abson and Roberts[15]	1946	Lead sulphide photo-conductive cells Carbon composition resistors	$b = 1{\cdot}62$ to 2·52 $c = 0{\cdot}8$ to 1·8 $b = 1{\cdot}7$ to 1·9 $c = 1$	 0·3 to 500 c/s 0·3 to 500 c/s
Schiff[16]	1946	Silicon crystals	$b = 2$	—
Miller[17]	1947	,, ,,	$c = 1$	50 c/s to 1 Mc/s
Ryder and Kircher[18]	1949	Germanium transistors	Function of the collector voltage	—
Chipman, Campbell,[5] Pavich[6]	1948 and 1949	Commercial resistors of various makes and compositions	$a = b = 2$ $c = 1$ at 20 kc/s 1·7 at 500 kc/s (Campbell) $c = 1$ (Pavich)*	 20 to 500 kc/s

The letters a, b, c, d refer to the indices in the expressions for the noise voltage given below:

$$e^2 = \tilde{k} R^a I^b f^{-c} \text{ mean square volts per unit bandwidth}$$
$$e^2 = k' R^a V^d f^{-c} \text{ mean square volts per unit bandwidth.}$$

Where R is the resistance value, I is the mean direct current flowing, V is the voltage applied across the resistor and f is the frequency of measurement.

* Pavich used an improved input circuit on his noise amplifier and with it obtained a constant value of unity for c. This was verified by the present writer.

produce a measurable noise voltage. In the absence of any previously recorded data, it seemed logical to assume that this magnitude, for low frequencies at least, would be similar to the magnitude of the direct voltage required to produce measurable noise in past experiments. This is of the order of tens or hundreds of volts, depending on the resistance and power rating of the resistor. With such voltages and a narrow measuring bandwidth, noise voltages in the region of one microvolt are produced.

The assumption of the exciting voltage requirements lays bare the first real difficulty—that of separating the noise from a signal differing in magnitude from it by some 140 db. This problem exists in the d.c. case, but there it is answered by the fact that there is always an infinite frequency ratio between the noise and the exciting signal. The a.c. case, however, must be solved either by extremely efficient filtering or by using a wide separation in frequency between the two signals. The emphasis should be laid on filtering, however, as it seems important to study the noise output at frequencies both near to and remote from the exciting frequency.

Trials of Methods

Several possible methods were discussed before any actual work was begun, and others presented themselves while the experiment was in progress. These are discussed below, more or less in the order in which they came up.

1. *Use of LC Circuits.* This includes the use of simple tuned circuits in the input of wideband amplifiers, tuned amplifiers, and more complex *LC* filter circuits.

The first test was made with the equipment which had been used the previous year for d.c. noise measurements.[6] This consisted of a tuned radio-frequency noise amplifier covering the frequency range from 16 to 600 kc/s which was isolated from the resistor under test by a two-stage wideband amplifier.

In the test, the resistor was excited by a 100-kc/s signal of 40-V r.m.s. amplitude provided from a signal generator and a converted radar video amplifier. A parallel *LC* circuit tuned to 100 kc/s was inserted in the input to the noise amplifier, as shown in Fig. 1. The tuned amplifier was set at 20 kc/s. This arrangement, with a coil having a Q of approximately 200, gave a rejection at 100 kc/s of about 60 db, but this was insufficient. The second stage of the wideband amplifier was blocked by the remainder of the exciting signal.

A second test, using the tuned amplifier alone with a high resistance in series with its input, was also unsuccessful.

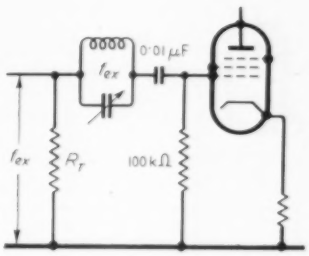


Fig. 1. *The use of a simple LC tuned circuit to reduce the level of the exciting signal (f_{ex} = exciting frequency).*

TABLE 2

THEORIES ON CURRENT NOISE

Bernamont	Variation in the number of free electrons based on random dissociation (and recombination) of atoms and electrons. The result gives $a = b = 2$ and allows the index c to take values between 0 and 2. Makes the noise inversely proportional to the volume of the conductor.
Brillouin[19]	A theoretical equation, from quantum statistics, for the total noise energy over the entire spectrum which explains Bernamont's results and shows how the index a can vary with current density.
Otto	Explained the linear dependence of noise on resistance as found in his experiments by the use of contact theory. He found the noise to be independent of volume, in contradiction to Bernamont. Suggested that noise is a property of semi-conductors in general.
Christensen and Pearson	Derived the frequency dependence by postulating two regions of conduction—one, stable, and carrying the majority of the current, and the other, unstable, due to thermal vibrations, causing the noise. Using analysis similar to Otto, they deduced possible values for the index a of between 1 and 3.
Schottky[20]	A theory explaining flicker noise in valves as due to the diffusion of impurity on to the cathode surface, where they affect the work function and emission temporarily.
Surdin[21]	Used a modified form of Bernamont's current noise expression to explain flicker noise in valves.
Davydov[22] and Gurevich	Explained the voltage fluctuations in semi-conductors by assuming a shot-effect mechanism in the barrier layers and, by applying a Fourier integral to a current pulse, derived a formula for the spectral density of the noise somewhat similar to Bernamont's.
Macfarlane[93]	Theoretical explanation of flicker noise somewhat similar to Schottky but allows the impurity atoms to occupy only a small part of the cathode surface and considered them to be acted on by electric conduction as well as diffusion. Extends the theory to cover all current-noise phenomena.

The use of more complicated *LC* filters was not attempted, partly because of the lack of readily-available components and partly because of the apparent unwieldiness of the method.

2. *Crystal Filters.* In this method, pairs of crystals would be required, one to stabilize the frequency of the generator and the other to act as the filter. This method was also not attempted for much the same reasons as above.

3. *Selective RC Circuits.* Circuits such as the parallel-T,[24] assembled from quite ordinary components and fitted with adequate trimmers, can give selectivities of the same order as crystals. This was another reason for the adoption of this method after the failure of the simple *LC* circuits.

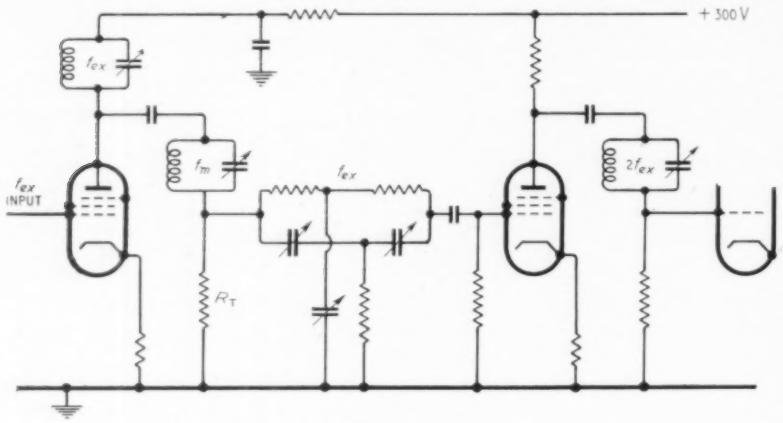


Fig. 2. *The first successful circuit using the parallel-T. R_T is the resistor under test, f_{ex} and $2f_{ex}$ are the exciting frequency and its second harmonic, respectively, and f_m is the frequency at which the noise is measured.*

The circuit of Fig. 1 was used again, with the T substituted for the parallel *LC* circuit. The frequencies used were the same. At first, the results were not encouraging, the maximum attenuation being only slightly greater than that obtained with the *LC* circuit. On examination of the output from the parallel-T, however, it was found that this consisted almost entirely of a 200-kc/s signal—the second harmonic of the exciting signal. When this harmonic was reduced by the insertion of a parallel *LC* circuit, tuned to 200 kc/s, between the two wideband stages, the attenuation was increased to between 90 and 100 db. This limit was found to be imposed by the presence of higher harmonics in the exciting signal but, as the amplifier did not become overloaded by this remainder, there was no need to proceed further with the filtering.

At this point, the exciter began to present difficulties for it was now possible to examine the output of the tuned amplifier for the presence of noise. The exciter was a surplus video amplifier with a bandwidth of some 3 Mc/s driven by a radio-frequency signal generator. It was found that the 20-kc/s level was greatly increased when the exciter amplifier was switched on, whether there was any input to it or not. This, of course, was due to noise generated in the exciter itself. A tuned circuit was fitted in the exciter output and tuned to the exciting frequency (100 kc/s). This effectively solved the problem, but it posed another since it lowered the exciter output impedance at the measuring frequency.

Since the noise resistor can be considered as an equivalent generator having an internal impedance approximately equivalent to its d.c. resistance then, for maximum transfer of noise power, the combined load impedance of the exciter output, the noise amplifier input and any calibrating device must be approximately equal in magnitude to the d.c. resistance of the resistor.

In order to raise the exciter output impedance at the measuring frequency, still another tuned circuit was inserted in series with the exciter output. Fig. 2 shows the essential parts of the circuit at this stage. With this addition, significant differences in the 20-kc/s level could be observed when resistors having different d.c. noise characteristics were placed in the test position.

The 20-kc/s signal was detected by means of a vacuum thermocouple, and the thermocouple output was fed through a chopped amplifier to a recording milliammeter. Several recordings were taken for two resistors, one a Welwyn pyrolytic and the other an I.R.C. carbon-composition type. Excitation levels of from 10 to 40 V r.m.s. were used at frequencies of 100 and 500 kc/s. As a form of calibration, the recordings include noise from d.c. excitation of the same resistors and some shot noise levels from a CV172 diode. Composite sketches from two of the best of these recordings appear in Fig. 3.

In these first tests, the measurements were always carried out at a frequency below the exciting frequency (viz., 20 kc/s), as this appeared to be the easier method of the two, there being no trouble from harmonics of the excitation. After the first success of the method, however, several attempts were made with the measuring

frequency above that of the excitation. It was found that, even with the measuring frequency lying between two harmonics of the excitation, the noise amplifier output increased whenever the excitation was switched on, regardless of the type of resistor in the test position. No amount of adjustment of the filter would alter this effect and, as there appeared to be no difference in the results obtained for the different types of resistors, the attempt was abandoned. It was thought to be more worth while to continue with the first method and so obtain some results than to spend a great deal of time on a doubtful course.

It must be noted that the observations in this case were all carried out by a visual observation of the noise amplifier output on an oscilloscope. The usual detector and recording device were never used. Much later, a question arose regarding this, with the suggestion that the rectified output might have shown a significant change for different types of resistors, but the equipment had been dismantled and time did not permit a re-examination of the method. There is, then, a possibility that results can be obtained with this method.

The experiment so far showed that there is noise produced by the flow of alternating current in some resistors and that it could be measured. The existing apparatus, however, was unwieldy in that it required several adjustments for each change of frequency. The exciter amplifier, also, was not satisfactory in that it did not cover a sufficiently wide frequency range and that its output voltage was not large enough to load the test resistor. It was decided that a new exciter, having wide frequency range, high output and low noise should be built. At the same time, it was thought worth while to examine one or two other possible methods.

4. *Balanced Input Circuits.* In this method, the exciting signal alone is fed to one input terminal and the excitation plus noise to the other. Two circuits using valves were tested, their essentials appear in Fig. 4.

In Fig. 4(a) the exciter is a cathode-loaded phase inverter. The balancing valves are low-μ triodes with heavy cathode feedback and input circuits as similar as possible. In one, the grid resistor is the test resistor, in the other, it is a 'noiseless' wire-wound or cracked-carbon type. Feedback is necessary to prevent the grids being driven off the linear portion of the characteristic by the large amplitude of the exciting signal. This arrangement gave considerable attenuation of the exciting signal, but

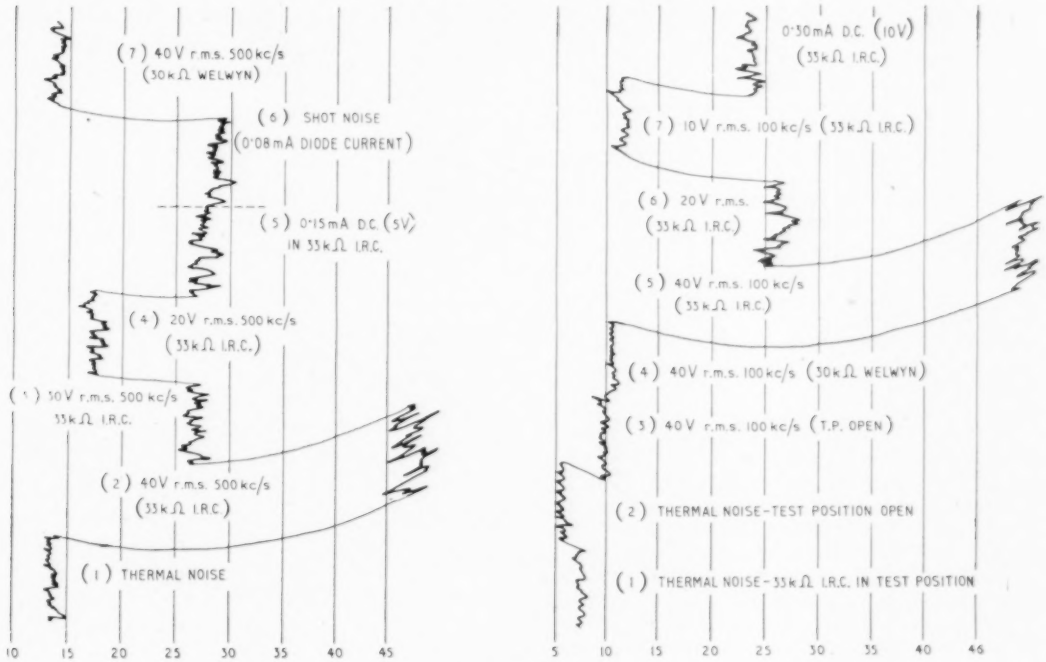

(7) 40 V r.m.s. 500 kc/s
(30 kΩ WELWYN)
(6) SHOT NOISE
(0·08 mA DIODE CURRENT)
(5) 0·15 mA D.C. (5V)
IN 33 kΩ I.R.C.
(4) 20 V r.m.s. 500 kc/s
(33 kΩ I.R.C.)
(3) 10 V r.m.s. 500 kc/s
33 kΩ I.R.C.
(2) 40 V r.m.s. 500 kc/s
(33 kΩ I.R.C.)
(1) THERMAL NOISE
10 15 20 25 30 35 40 45
0·30 mA D.C. (10V)
(33 kΩ I.R.C.)
(7) 10 V r.m.s. 100 kc/s (33 kΩ I.R.C.)
(6) 20 V r.m.s.
(33 kΩ I.R.C.)
(5) 40 V r.m.s. 100 kc/s
(33 kΩ I.R.C.)
(4) 40 V r.m.s. 100 kc/s (30 kΩ WELWYN)
(3) 40 V r.m.s. 100 kc/s (T.P. OPEN)
(2) THERMAL NOISE-TEST POSITION OPEN
(1) THERMAL NOISE-33 kΩ I.R.C. IN TEST POSITION
5 10 15 20 25 30 35 40 45

Fig. 3. *Recordings of output of 20-kc/s noise amplifier with the circuit of Fig. 2.*

tended to amplify the even harmonics and was thus unsuitable.

The circuit of Fig. 4(b) gave higher attenuation, of the order of 60 or 70 db, but again this was inadequate. Grid and cathode of the mixer were fed in phase, with the noise being applied to the grid and the pure excitation to the cathode. The limit of attenuation was apparently set by the amount of signal coupled directly through the interelectrode capacitances of the valve. There was no marked amplification of harmonics in this case as the effect of curvature of the adequate filtering. This figure is easy to obtain with a few stages of *RC* filtering and such a filter has the advantages of being easy to build and maintain. The problem of the exciter output impedance may be solved by the use of a high-pass filter which will pass the exciting signal without great attenuation but which prevents the noise from reaching the low impedance of the output circuit. In both cases, however, if *RC* filters are used, the exciting and measuring frequencies must be separated by at least a factor of ten in order that the filters can do their work.

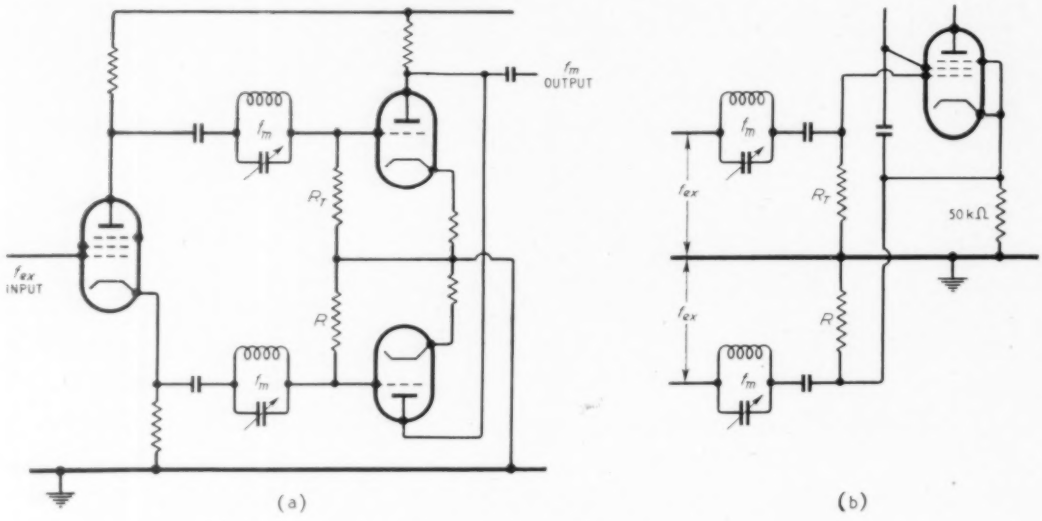


Fig. 4. *Two forms of balanced input which were tried without success. R_T is the test resistor and R is a wire-wound resistor of the same value.*

characteristic was cancelled by having the grid and cathode rise and fall together.

Another method of balancing might be arranged by using a transformer with a double primary as the input element. Here the difficulty of getting an accurate balance, preventing capacitive coupling and suppressing harmonics would probably cause a great deal of trouble.

5. *Combinations of Highly-Selective Circuits with RC High- and Low-Pass Filters.* By using a selective circuit such as the parallel-T in conjunction with a low-pass filter, the exciting signal can be removed almost completely. This method is only useful where the measuring frequency is below that of the excitation. The parallel-T is tuned to the fundamental exciting frequency and the low-pass filter eliminates the harmonics.

Where the ratio of the two frequencies is ten or more, the low-pass filter need only attenuate by about 20 db at the second harmonic to give

Description of Apparatus

The following is a fairly detailed description of the apparatus as it was used for the bulk of the measurements. At this stage in the work, a high-gain audio-frequency amplifier was made available. This unit had many desirable features which made it very useful as a noise amplifier. These were: a gain of 90 to 100 db, variable bandwidth by the use of filters which could be switched in, anti-vibration mountings on the first two stages and, in a separate chassis, a stabilized high-voltage supply and a rectifier to supply the heaters with direct current. This amplifier became the central unit in the final apparatus.

Fig. 5 is a block diagram. On the top row, from left to right, it shows the radio-frequency signal generator—the source of the exciting signal. Next is the radio-frequency amplifier or exciter, followed by the high-pass filter and the test

position where the test resistor, calibrator and voltmeter probe were situated. These three units and the low-pass filter are shown in a dotted box which is intended to indicate that they were all on the exciter chassis. The next box represents the noise amplifier with the parallel-T filter at its input. These are followed by the detector and indicating device. The blocks below show the dispositions of the power supplies and the division of the load between the two constant-voltage transformers. In this group, one block represents the valve voltmeter (Hewlett-Packard Model 410A) which was used to determine the excitation level at the test resistor.

The Exciter Chassis. Fig. 6 is a circuit diagram of the complete exciter chassis. The exciter amplifier is a three-stage, low-power amplifier. The first two stages were designed to cover a frequency band between 20 kc/s and 7 Mc/s at the half-power points. The response below 20 kc/s is made low by the addition of high-pass filtering between stages to prevent the amplification of noise in the band to be examined. The final stage has a tuned anode load which improves the purity of the output and keeps the output up at high frequencies. This tuned load consists of a mica trimmer capacitor with a range of 600–1,500 pF and a set of plug-in coils. The coils used were borrowed from a set provided with a Boonton 'Q' Meter. They were chosen by trial and error to give a required set of frequencies, and the trimmer capacitor, which gave a 20 to 30% variation in frequency, was used for final adjustment. This amplifier gave signals of up to 100 V r.m.s. at the test position at frequencies up to 5 Mc/s when loaded with a 10-kΩ test resistor.

The High-Pass Filter. The exciter was isolated from the test resistor by the inclusion of a high-pass filter. This was a six-stage *RC* filter, designed to give approximately 3-db attenuation per stage at 5 kc/s. The shunt elements were all 200-kΩ resistors, but the series elements were graded so that the impedance of each was 1·4 times that of the one preceding it. The first series element, nearest the exciter, had an impedance of 3,000 Ω at 5 kc/s. With this arrangement, at 5 kc/s, the impedance seen from the test position, looking towards the exciter, was slightly above 20 kΩ.

The Test Position. The high-pass filter terminated in a pair of jacks where the test resistors were plugged in. One of these jacks was earthed directly to the chassis and the other was mounted in a polystyrene grommet. The test resistors were mounted on strips of polystyrene carrying two small banana plugs which fitted the jacks. The insulated jack was connected to the live side of the filter, to the calibrator diode anode and to a separate pin jack into which the valve voltmeter probe was fitted during excitation-level measurements. On the output side, the insulated jack was connected to a low-pass filter.

The Low-Pass Filter. This was a simple *RC* filter consisting of five similar stages of 10-kΩ series elements and 300-pF shunt elements. The output of this filter was connected to the noise amplifier through a short length of coaxial cable.

The heater and anode-current controls and the anode-load resistor for the calibrator diode were also mounted on this chassis. In an effort to reduce drift in the calibrator anode current during readings, the heater controls were switched on to a dummy load of 4 Ω when the calibrator was not

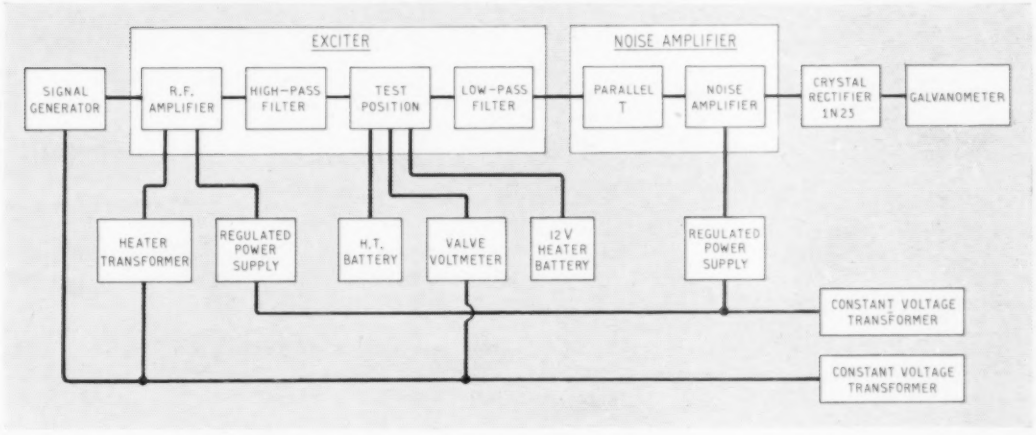


Fig. 5. *Block diagram of complete final equipment. The top row shows, from left to right, the excitation source, the exciter amplifier, filters and test position, noise amplifier, detector and indicator. Below these are the various power supplies and the valve voltmeter for excitation level measurements and the constant-voltage transformers for mains stabilization.*

in use. This kept the controls at or near the operating temperature and held the diode emission fairly constant when it was required.

The Noise Amplifier. This unit has already been mentioned at the beginning of this section. With some modifications, and operated in its narrowest bandwidth condition—approximately 2 to 5·5 kc/s—this amplifier became the core of the final equipment. Fig. 7 gives a response curve for the amplifier in this condition.

The modifications to the noise amplifier were as follows: The input transformer was removed and the first stage converted to resistance-capacitance coupling. The parallel-T filter was built into the space vacated by the transformer. The output was brought out through a switch so that it could be earthed through a 600-Ω resistor during changes and voltage measurements at the test position. This was necessary to protect the detector and galvanometer from surges. The whole unit was enclosed in a heavy sheet-steel box when it was found that the iron-cored chokes in the bandpass filters were subject to power-frequency pick-up. The gain control was altered from a continuously-variable potentiometer to a step control, of the type shown in Fig. 8. This allowed for fairly accurate resetting of gain and prevented the gain from shifting during a measurement due to jarring or inadvertent movement of the control.

The parallel-T filter was assembled in the space left vacant by the removal of the input transformer. It consisted of three fixed resistors and three small trimmer capacitors (3–15 pF approximately). These allowed tuning of the unit from just above 100 kc/s to above 500 kc/s. The components were assembled around an octal valveholder so that three 28-pF fixed capacitors could be plugged in to extend the range to 50 kc/s. For exciting frequencies above 500 kc/s the parallel-T was unnecessary as these high frequencies were very highly attenuated by the low-pass filter. It was left in position for all the measurements, however, as it formed a constant part of the load on the test resistor.

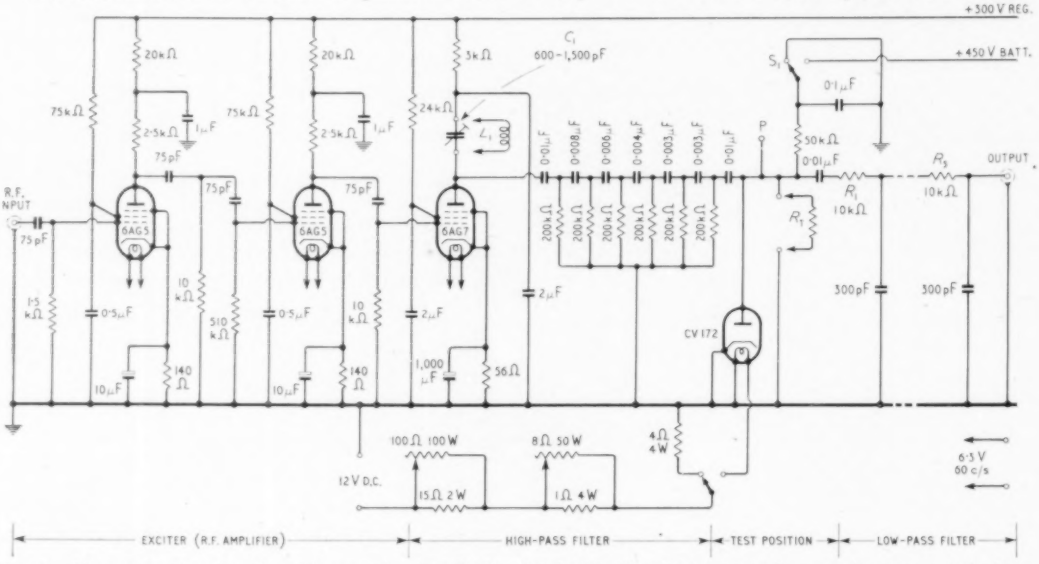


Fig. 6. *The exciter chassis:—L_1, 'Q' meter plug-in coil; C_1, 600-1,500 pF mica trimmer; S_1, ceramic rotary switch; R_1 to R_5, 10 kΩ 2 W Welwyn 5-sections; R_T, test resistor. All filter section condensers 500 V mica; all filter section resistors Welwyn or I.R.C. Nobelloy. 12 V d.c. supply through external control and limiting resistors.*

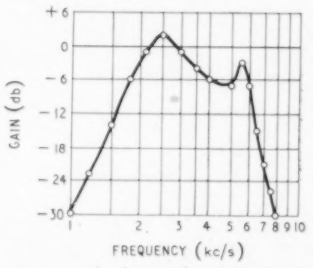


Fig. 7. *Noise amplifier response curve; 0 db ≈ 90 db gain.*

Detector and Indicator. The output of the noise amplifier was rectified by a 1N23 crystal diode built into the coaxial output line and the rectified output was used to drive a Leeds and Northrup galvanometer having a deflection sensitivity of 0·8 μV per millimetre on a scale at 1 metre from the meter movement. The output of the crystal rectifier was of the order of $\frac{1}{10}$ V and thus allowed

the use of a large resistance in series with the meter movement. This resistance in series with the meter impedance and in conjunction with the damping of the movement provided adequate smoothing of the rapid fluctuations in the signal.

At the beginning of the tests with this equipment, a low-pass filter consisting of two sections of 400 μF and 2·5 kΩ each was used for smoothing the voltage applied to the galvanometer. This held the meter quite steady for the greater part of the time, but occasional large peaks of short duration tended to drive the meter completely off scale with such force that it generally became locked on the end stop. With the capacitors removed, the general action of the meter was much more lively, but the extreme overshoots were eliminated.

Since the voltage applied to the galvanometer is subject to random variations, it is necessary that some method of reading should be used which will provide a time average of the deflections. The previous observer on this project achieved this by watching the light spot in its movements for 10 minutes or more for each reading and arriving at the average deflection mentally. The present writer, in his first month on the project, was engaged in checking some of the earlier results and, during this time, found that, apart from the great concentration required and the strain produced by this method, he was not able to decide on an average deflection that was reproducible to better than 10%. He then decided to take a series of instantaneous readings at fixed intervals of time and to find the average of these. This method, by taking 60 readings at 5-second intervals, produced results which varied one from another by 5% or less. A more rapid method was developed later in which the galvanometer scale was illuminated for a fraction of a second every 2 seconds. The deflections revealed by these flashes were recorded on an electro-mechanical calculator which added them and also showed the number of readings and the average.

Power Supplies. The exciter amplifier anode voltage was supplied by a regulated laboratory unit and the heaters from a separate transformer. The noise amplifier supply has been mentioned previously, but a further note must be added regarding it. During the early tests, it was found that the gain was varying with line voltage and this was found to be due to the poor regulation of the d.c. heater supply for the noise amplifier. This variation was eliminated by the use of constant-voltage transformers in the 60-c/s supply. Two 250-VA transformers were used and the load was divided between them as shown on the block diagram.

The calibrator diode was supplied from a bank of ten 45-V dry batteries for the anode and a 12-V accumulator for the heater. This heater battery was kept on charge whenever the calibrator was not in use. Filtering, metering and controls on these two supplies were straightforward.

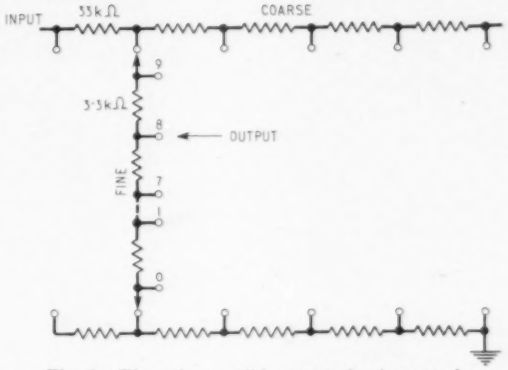


Fig. 8. *The noise amplifier stepped gain control.*

Faults in the Equipment

It has been noted in the description of the equipment that several faults appeared during the preliminary tests. These all required considerable time to detect and correct, but one, which has not yet been mentioned, caused more trouble and took more time to clear than all the rest.

When the equipment was first assembled there were no resistors of the deposited-carbon or metal-film type available for use in the filters. According to the results obtained by Pavich the next best type with regard to current-noise output was the Allen-Bradley composition type. These were available and were used in the filters. For some weeks, while the performance was being tested and faults cleared, they performed adequately. Then, large increases in the noise level began to appear and, as time went on, the frequency of appearance of these bursts increased. When they first appeared the bursts were thought to be due to electrical disturbances originating elsewhere in the laboratory, but this was disproved after many observations.

Eventually, it was discovered that the first resistor in the low-pass filter had become noisy. When it was placed in the test position it was found to be producing approximately three times as much noise as the test resistor. This represented an increase in noise level of approximately 40 db. It was replaced by a Welwyn Pyrolytic type resistor and the fault disappeared. Later, the bursts appeared again and were found to be originating in the second section of the low-pass filter. Shortly after this, a stock of Welwyn resistors arrived and both filters were fitted with these. The fault did not reappear.

(*To be concluded*)

PULSE FREQUENCY MULTIPLICATION AND DIVISION BY DELAY LINES

By Dr. techn. Heinz Zemanek

(Institute of Telecommunication, Technische Hochschule, Vienna, Austria)

Pulse Multiplication

IT is a special feature of pulse techniques that pulse-repetition frequencies can be multiplied by linear devices, for example, by delay lines. A series of periodic pulses of relatively short duration contains all necessary harmonics, so that by adequate suppression and attenuation any multiple of the original pulse frequency may be generated. The delay line, matched at one end by Z_0, has tapping points after every i section and the voltages at them are fed through resistances R into one common resistance R_c (Fig. 1). If there are k resistances R, the output is a pulse series with a frequency k times as high as that of the input.

Assuming rectangular pulses of duration t_p, amplitude V_1 and interval $T = 1/f$ then, if the influence of attenuation and phase distortion is neglected, the original train

$$v_1 = \frac{t_p}{T} V_1 \left\{ 1 + 2 \sum_{n=1}^{\infty} \frac{\sin n\pi \frac{t_p}{T}}{n\pi \frac{t_p}{T}} \cdot \cos 2\pi nft \right\}$$

will generate across the resistance R_c the output voltage

$$v_2 = \frac{t_p}{T/k} V_2 \left\{ 1 + 2 \sum_{n=1}^{\infty} \frac{\sin nk\pi \frac{t_p}{T}}{nk\pi \frac{t_p}{T}} \cos 2\pi nkft \right\}$$

Only the multiples of the frequencies $k.f$ appear, and all components $n.f$ with $n \neq m.k$ (with $m = 1, 2, 3 \ldots$) are suppressed. Of course, the passive network conserves the energy, and the multiplication of pulses is accompanied by a diminution of the amplitudes. Assuming that the input is also matched (see Fig. 2), at the first moment $V_1 = V_0/2$. If all the pulses of a group arrived simultaneously at R_c (Fig. 2), we should get for the voltage ratio

$$\frac{V_2}{V_1} = \frac{V_2}{V_0/2} = \frac{2R_c}{Z_0 + 2R_c + 2R/k}$$

The corresponding powers would be

$$\frac{N_2}{N_1} = \frac{4R_c \cdot Z_0}{(Z_0 + 2R_c + 2R/k)^2} < \frac{1}{2(1 + 2R/kZ_0)} < \frac{1}{6}$$

But since R_c receives this total energy in k parts as k outcoming pulses, the energy of one outcoming pulse is less than $1/6k$ of the energy of an incoming pulse. This expression gives the maximum power that can be delivered to R_c.

MS accepted by the Editor, November 1953

Pulse Division

Unlike multiplication, division requires the use of a non-linear device, because new frequencies, lower than the pulse frequency of the incoming series, and all their multiples must be generated. With a pentode as a non-linear element, the delay line can be used for division as shown in Fig. 3.

A positive pulse v_1 on the third grid opens the pentode, so that a negative pulse appears on the anode, comes to the output and also enters the delay line. At the moment when the next pulse v_1 might open the valve, the negative one from the resistance R_1 is blocking the cathode current on grid 1 and no pulse comes to the output and into the line. As the negative pulse travels along the line, it always prevents current at the moments when the pulses of v_1 would otherwise allow it. Only when the propagating pulse has passed the last resistance R_{k-1} can the next positive pulse v_1

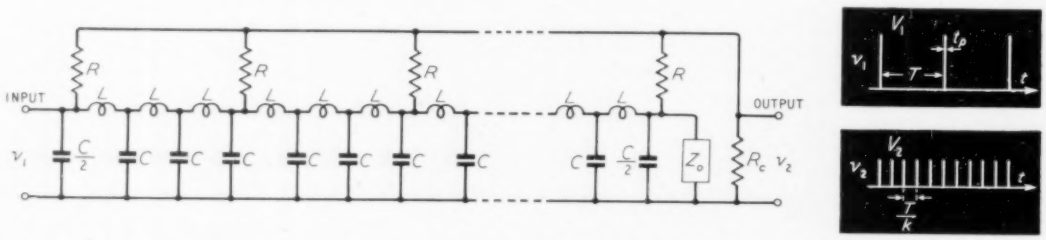


Fig. 1. *Frequency multiplication by delay line. To avoid loss of amplitude the combined resistance of all the resistances R in parallel should exceed Z_0.*

begin another period, k times as long as the period of the incoming pulse series v_1; only one of each k pulses arrives at the anode and the output.

Instead of a pulse series, the input can be sinusoidal, every positive part operating then as a pulse. The output consists of pulses with a duration of $1/2f$ at intervals of $k/f = kT$.

Both circuits work only for the frequency for which they are designed. But this frequency is correctly multiplied or divided; not a single pulse is lost. As the division is achieved by a kind of feedback, under certain bias conditions the output, once originated, is maintained, even if the input stops. In this case the frequency is determined by the delay of the line. As soon as the input begins again, correct division is immediately resumed. With a certain other bias, the output disappears with the last pulse of the input.

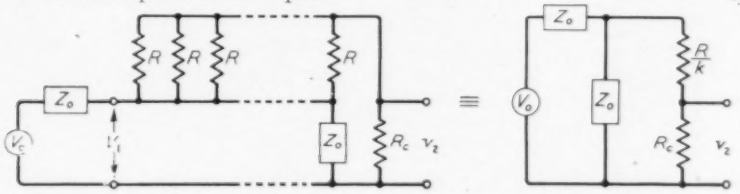


Fig. 2. *D.C. equivalent circuit.*

The pulse division requires enough voltage to block the pentode; therefore Z_0 must not be chosen too low, since $V_2 = g_m Z_0 V_3$.

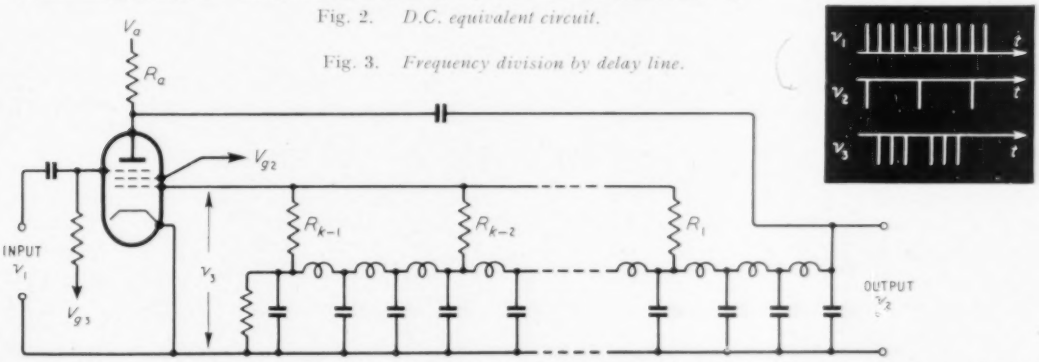


Fig. 3. *Frequency division by delay line.*

The Delay Line

There are plenty of publications on delay lines;* we will give here only a very short discussion of the line with lumped elements, sufficient for simple purposes of frequency multiplication and division. The line may be considered as a low-pass filter (Figs. 1 and 3) with

$$\text{Characteristic impedance } Z_0 = \sqrt{\frac{L}{C}};$$

$$\text{Cut-off frequency } f_c = \frac{1}{\pi\sqrt{LC}};$$

$$\text{Delay time per coil } \tau_0 = \sqrt{LC}.$$

The shortest pulse form which propagates correctly along a line with a cut-off frequency f_c has the duration of a Nyquist interval $t_0 = 1/2f_c$. For the duration to be doubled, i.e., for

$$t = 2\,t_0 = \pi\sqrt{LC} \approx 3\sqrt{LC} = 3\,\tau_0$$

three sections between two tapping points are needed ($i = 3$). Therefore a division by k requires $3\,(k-1)$ coils, and a multiplication by k at least the same number. When a certain pulse form must be maintained in the system, this form determines the cut-off frequency and thereby the number of coils. The number of capacitors of capacitance C is one less, but two of capacitance $C/2$ are at the ends. If Z_0 and f (the output frequency of multiplication or the input frequency of division) are given, the values of L and C are calculated as follows:

$$L = \frac{Z_0}{3f}; \quad C = \frac{1}{3fZ_0}$$

The losses in the coils cause an attenuation of the propagating pulse, which can be equalized by varying the values of the resistances R. But with increasing numbers of coils, the amplitude is weakened too much for sure operation; the limit of working depends very much on the quality of the coils.

***Literature on Delay Lines**

M. J. E. Golay, "The Ideal Low-Pass Filter in the Form of Dispersionless Lag Line", *Proc. Inst. Radio Engrs*, March 1946, Vol. 34, p. 138 P.

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MULTIPLEX TELEVISION TRANSMISSION

Sub-Carrier and Dot-Interlace Systems

By J. Haantjes and K. Teer

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(Concluded from page 233, September issue)

5. Signal Separation at the Receiver

With both methods of transmitting several signals in a single channel the separation of the signals at the receiver must be effected by a process known as 'synchronous detection'. The modulated sub-carrier system requires the presence in the receiver of a waveform $\cos \omega_p t$ and the dot-interlace system needs

$$P_k = 1 \pm 2 \cos (\omega_p t + \phi_k) + \cos \{2(\omega_p t + \phi_k)\} \pm \ldots .$$

These separating waveforms must be available at the receiver with the right frequency, phase and shape. The information about frequency and phase has to be conveyed to the receiver by an extra synchronizing signal and one example of this is the 'burst' introduced into the back porch of a line sync pulse in the American colour system,[3,4] This is shown in Fig. 15.

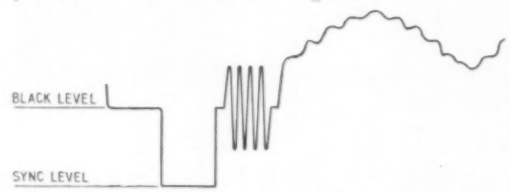


Fig. 15. *The position of the 'burst' in the N.T.S.C. colour-television signal.*

If synchronous detection is used, therefore, the receiver must contain devices for generating these synchronized separating waveforms and for multiplying the received signal by them. This necessarily involves the use of rather complicated circuitry at the receiver, which must be considered a disadvantage of this method of signal separation. In the rest of this article, therefore, the possibility of simplifying the separation circuits of the receiver will be considered.

First of all, however, a possible simplification of synchronous detection should be mentioned. If the sub-carrier is always present with the reference phase in the transmission it can be filtered out and, after amplification, it can act either as a separating waveform itself or as the input to a separating waveform generator. For instance, when the transmitted signal is

$$S_1 \pm (S_2 + k) \cos \omega_p t$$

where S_2 varies between 0 and S_{2max} and k is a constant >0, the sub-carrier is present in the right phase for all possible values of S_2, as shown in Fig. 16(b).

However, this arrangement has the disadvantage that the peak value of the signal containing the S_2 information, namely $S_2 + k$, is much larger than is necessary for obtaining a given signal-to-noise ratio at the receiver.

When the detection is synchronized by a 'burst' the transmission signal can, with advantage, be chosen to be

$$S_1 + (S_2 - \tfrac{1}{2}S_{2max}) \cos \omega t$$

in order to obtain the same S_2 signal. As shown in Fig. 16(c) the peak value of the modulated sub-carrier is then only $\frac{1}{2}S_{2max}$.

In general, an unnecessary increase in the peak value of the transmission signal is undesirable. An increase in the peak value of the cancelling signal is especially undesirable in a transmission system which is intended to be compatible. The cost of simplifying the receiver circuitry, therefore, is an

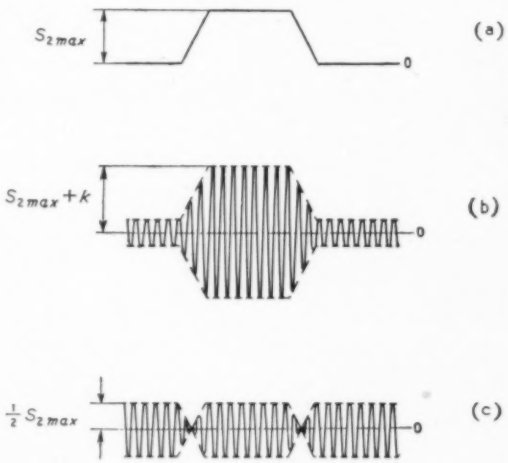


Fig. 16. *The signal S_2 is shown at (a), $(S_2 + k) \cos \omega_p t$ at (b) and $(S_2 - \frac{1}{2}S_{2max}) \cos \omega_p t$ at (c).*

increase of interference in normal reception, resulting from the increased peak value of the cancelling signal. It is, however, not without importance, for the resulting waveform of the transmission signal permits the use of non-synchronous detection.

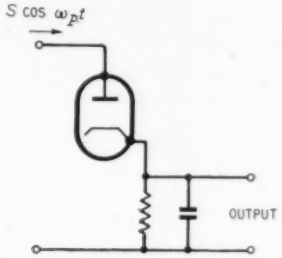


Fig. 17. *A simple detection circuit.*

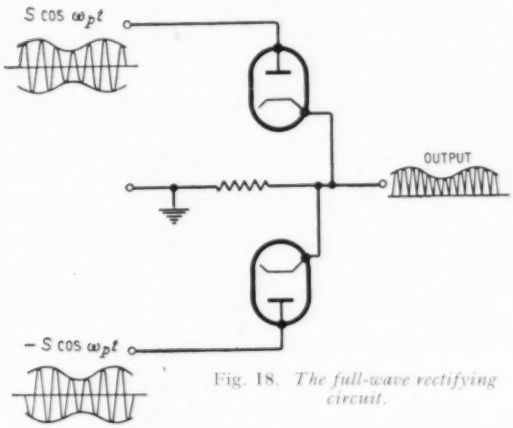


Fig. 18. *The full-wave rectifying circuit.*

6. Non-Synchronous Detection

6.1. *Conventional Detection of a Sub-Carrier*

In the search for simpler separation methods the first step is to try to use ordinary detection of the sub-carrier with a non-linear circuit element of the form shown in Fig. 17. It is well known that a modulated carrier $(S_2 + k) \cos \omega t$, where S_2 varies between 0 and S_{2max}, and k is a constant, can be demodulated by such an element if $k > 0$. A sub-carrier with a peak value of at least S_{2max} can, therefore, be detected in this way, whereas the signal of Fig. 16(c) with a minimized peak value cannot. There is a further restriction imposed by the bandwidth of S_2. Detection can be carried out with the circuit of Fig. 17 only when the bandwidth is small compared to f_p. When it is not, this circuit must be replaced by a full-wave rectifier, such as that of Fig. 18, which is able to demodulate frequencies up to f_p.

It is, however, also possible to demodulate the sub-carrier in the i.f. stage. The sub-carrier frequency is there transformed to $f_h + f_p$ (f_h being the vision carrier) and therefore large compared with the bandwidth of S_2. The vision carrier has, of course, to be suppressed before detection.

No account has so far been taken of the signal S_1 which is also present in the same band. As well as demodulating the sub-carrier, the detection circuit should also transform the interfering signal S_1 into cancelling signal components. As the S_1 frequencies have the cancelling relationship described in Section 2 this condition is satisfied if the detector is able to interpret them as sideband frequencies of f_p. It is obvious that this is impossible if, in some interval, the sub-carrier amplitude is only small or even entirely absent. For instance, when $S_2 = 0$ in $S_2 \cos \omega_p t$ there is no sub-carrier and the transformed S_1 interference signal in the detector output will have the same shape in successive pictures (apart from any change in the S_1 information within one picture period).

Because of this, in order to ensure the presence of a cancelling S_1 interference in S_2 the sub-carrier, amplitude must never fall below a certain level which is fixed by the peak amplitude of the S_1 interference. The transmission signal, therefore, must be $S_1 \pm (S_2 + k) \cos \omega_p t$

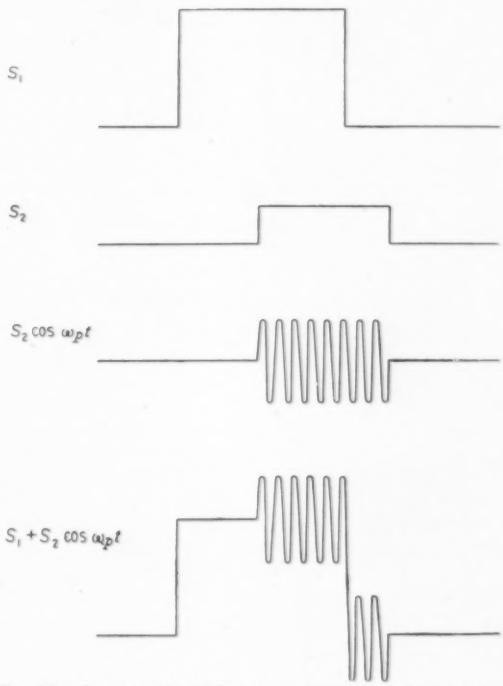


Fig. 19. *An example of the transmission signal of a sub-carrier system where the sub-carrier is* 100% *modulated.*

where in this case k has to exceed a certain value, k_1 depending on the peak amplitude of the interfering S_1 signal in the detector input.

The peak value of the modulated sub-carrier, already greater than the minimum possible value in order to make conventional detection possible, must therefore be increased once again to ensure the cancellation of the cross-talk in S_2 by S_1.

As it is important to keep this value small, another method of avoiding the non-cancellation of S_1 interference has been investigated. The method makes use of a special detection circuit and experiment has shown it to function with but little interference. The transmitted signal is

$$S_1 \pm S_2 \cos \omega_p t$$

and S_1 and S_2 vary between 0 and S_{1max} and S_{2max} respectively. An example of it is shown in Fig. 19.

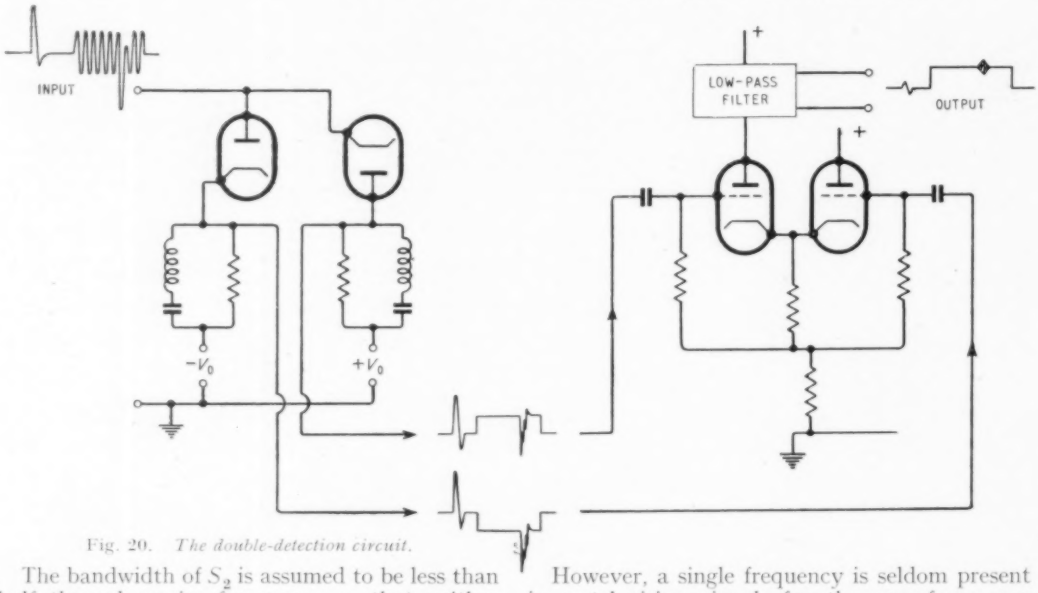


Fig. 20. *The double-detection circuit.*

The bandwidth of S_2 is assumed to be less than half the sub-carrier frequency, so that neither detection in the i.f. stages nor full-wave detection is necessary. The transmission signal is fed to a band-pass filter and its output is detected by the two detectors of Fig. 20, of which one follows the upper envelope of the signal and the other the lower envelope. As the upper envelope can reach below, and the lower envelope can reach above, the zero-carrier level, due to the S_1 signal, the diodes have to be biased by $+V_0$ and $-V_0$ respectively. The output signals of the two detectors have the carrier envelopes in opposite phase, but the S_1 interference is partly in the same phase. The signals are, therefore, subtracted and in Fig. 20 this process is carried out in a double-triode with common cathode resistances. The resulting output is taken to a low-pass filter of bandwidth S_2.

Compared with the conventional single-detector circuit this arrangement showed experimentally a considerable reduction of non-cancelling interference. The circuit does not alter the amount of cancelling interference.

The proper functioning of the circuit depends upon the shape of the interfering S_1 signal in the detector input. For example, when in some interval this signal is a single frequency near the sub-carrier and the sub-carrier itself is absent ($S_2 = 0$), this S_1 signal will be detected as if it were a carrier, as shown in Fig. 21. Under these conditions the detectors cannot follow the S_1 signal waveform, nor can they develop equal outputs to neutralize each other by subtraction.

However, a single frequency is seldom present in a television signal, for the waveforms are usually of step-function type, such as those of Fig. 19.

In order to produce a satisfactory picture from the S_2 information the double-detector circuit of Fig. 20 needs a peak value for the sub-carrier which is considerably less than the peak value necessary for the ordinary detector circuit. A further advantage is that it permits the use of simpler band-pass and low-pass filters, for frequencies normally suppressed by these filters are now neutralized by subtraction. The slope of the cut-off characteristic of the filters can be flatter and it is, in fact, undesirable that the cut-off should be very sharp. It is even possible to

omit the filters altogether, but the undesired signals in the output then become very large and the subtraction process is very critical.

6.2. *Non-Synchronous Separation for Two-Signal Dot-Interlace*

The possibility of using non-synchronous detection for the separation of the signals of a dot-interlace system will now be considered. As described in Section 4.5, the transmission signal for two signals can be written in the simplified form

$$S_1 + S_2 \pm \Omega (S_1 - S_2)$$

and the question arises as to whether the single-sideband sub-carrier which is present in this signal can be demodulated by a conventional detector.

From the discussion given in Section 6.1 it will be clear that the transmission signal has to be modified by the addition of a certain amount of sub-carrier. In the first place, the signal $S_1 - S_2$ can have negative values and so does not suit the operating conditions of a normal detector. Secondly, the addition of sub-carrier is necessary to ensure the cancellation of the unwanted signal $S_1 + S_2$. The double-detection circuit cannot be used here, because it can only detect a signal with a relatively small bandwidth. With dot-interlace the bandwidths of S_1 and S_2 are supposed to be equal to f_p and so the circuit cannot be used.

With sub-carrier added the transmission signal becomes

$$S_1 + S_2 \pm \Omega (S_1 - S_2 + S_{2max} + k)$$

where k has the same significance as in Section 6.1. The signal can now be detected by a full-wave rectifier, either in the i.f. or in the video stage.

I.F. Detection

In the i.f. stages the transmission signal appears as a modulation on the vision carrier of frequency f_h. Two carriers f_h and $f_h + f_p$ are, therefore, present and modulated by $S_1 + S_2$ and $S_1 - S_2$ respectively. Each carrier can be detected after the other has been suppressed and each detected output signal, $S_1 + S_2$ as well as $S_1 - S_2$, suffers interference from the sidebands of the unwanted carrier.

The information corresponding to S_1 and S_2 can be obtained from the sum and difference of the two detector output signals but, in consequence of the suppression of the undesired carriers, the result will show some attenuation of the highest frequencies of S_1 and S_2.

Video Detection

When the information is to be separated in the video stage it is not advisable to demodulate the sub-carrier by a full-wave detector in the way described above. The signals can be separated more simply if the transmission signal is again altered somewhat.

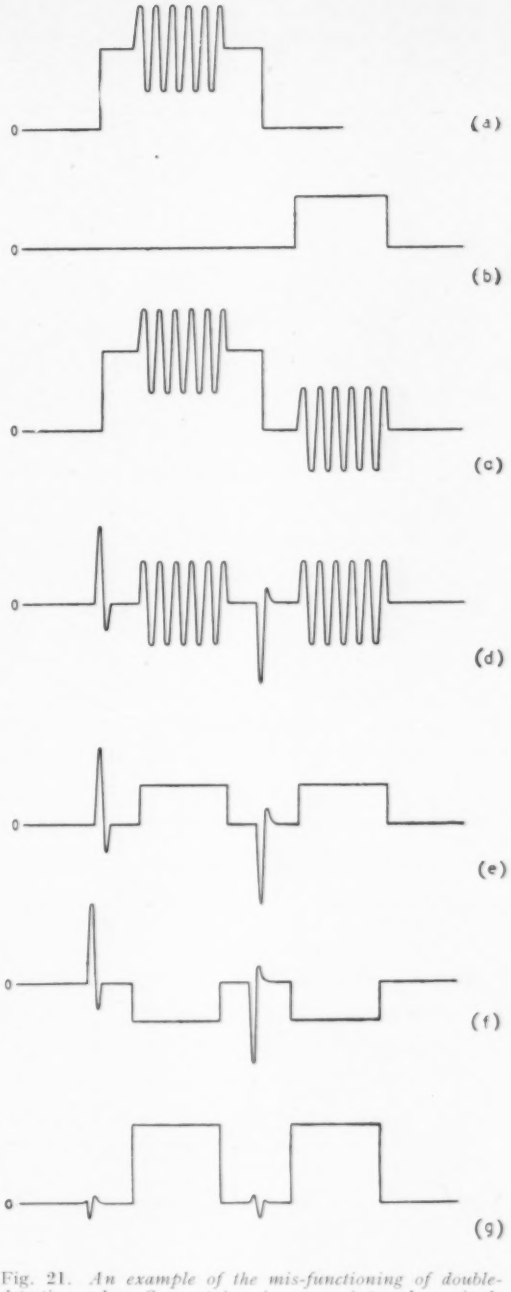


Fig. 21. *An example of the mis-functioning of double-detection when S_1 contains in some interval a single frequency near the sub-carrier frequency: (a) signal S_1; (b) signal S_2; (c) transmission signal; (d) input signal of the double detection circuit; (e) upper envelope; (f) lower envelope; (g) output signal of the double-detection circuit.*

In order to explain the principle of this method of separation, it is necessary to consider the shape of the signals occurring in a two-signal dot-interlace system. These signals S_1 and S_2 are supposed to vary between 0 and S_{max}. The signal S_1 is multiplied by

$$P_1 = 1 \pm 2 \cos \omega_p t + 2 \cos 2\omega_p t$$

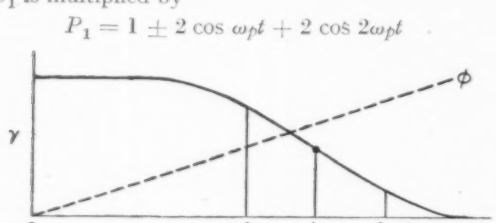


Fig. 22. *The transmission filter of a dot-interlace system for two signals.*

and then fed to a filter which has the characteristics of Fig. 22. It follows from the analysis of the dot-interlace system that the output signal is periodically equal to $2S_1$ with a repetition frequency of f_p. Midway between these moments the output signal is zero.

If there is no delay in the filter, and the + sign holds in P_1, the output signal equals $2S_1$ whenever

$$\omega_p t = 0 \pm 2n\pi$$

and is zero whenever

$$\omega_p t = \pi \pm 2n\pi$$

This can be verified by calculating the filter response for the input signal $S_1 = 1 + \cos \omega t$. The input signal of the filter is

$$\begin{aligned} S_1 P_1 &= (1 + \cos \omega t)\, P_1 \\ &= 1 + \cos \omega t + 2 \cos \omega_p t + 2 \cos \omega t \cos \omega_p t \\ &\quad + 2 \cos 2\omega_p t + 2 \cos \omega t \cos 2\omega_p t \\ &= 1 + \cos \omega t + \cos (\omega_p - \omega)\, t + 2 \cos \omega_p t \\ &\quad + \cos (\omega_p + \omega)\, t + \cos (2\omega_p - \omega)\, t \\ &\quad + 2 \cos 2\omega_p t + \cos (2\omega_p + \omega)\, t \end{aligned}$$

The output signal is, therefore,

$$\begin{aligned} &1 + \gamma_\omega \cos \omega t + \gamma_{\omega_p - \omega} \cos (\omega_p - \omega)\, t + \cos \omega_p t \\ &\quad + \gamma_{\omega_p + \omega} \cos (\omega_p + \omega)\, t \\ &\quad + \gamma_{2\omega_p - \omega} \cos 2 (\omega_p - \omega)\, t \end{aligned}$$

Because of the symmetry of the filter characteristic the relation $\gamma_\omega = 1 - \gamma_{2\omega_p - \omega}$ holds and so $\gamma_{\omega_p - \omega} = 1 - \gamma_{\omega_p + \omega}$. Hence, whenever $t = 0 + 2n\pi$,

$$1 + \gamma_\omega \cos \omega t + \gamma_{\omega_p - \omega} \cos \omega t + 1 + \gamma_{\omega_p + \omega} \cos \omega t + \gamma_{2\omega_p - \omega} \cos \omega t = 2 + 2 \cos \omega t = 2S_1$$

and whenever $t = \pi + 2n\pi$

$$1 + \gamma_\omega \cos \omega t - \gamma_{\omega_p - \omega} \cos \omega t - 1 - \gamma_{\omega_p + \omega} \cos \omega t + \gamma_{2\omega_p - \omega} \cos \omega t = 0$$

It is obvious that the positions of the $2S_1$ values and the zeros interchange for the pictures for which the negative sign holds.

In general, the output has the form shown in Fig. 23(a). When the higher frequencies in S_1 are of small amplitude the negative excursions of the output signal are almost negligible.

The second video signal S_2 is multiplied by

$$P_2 = 1 \mp 2 \cos \omega_p t + 2 \cos 2\omega_p t$$

with results similar to those obtained for the S_1 signal and illustrated in Fig. 23(b). As can be seen, the zeros of (b) coincide with the $2S_1$ values of (a), and the zeros of (a) coincide with the $2S_2$ values of (b).

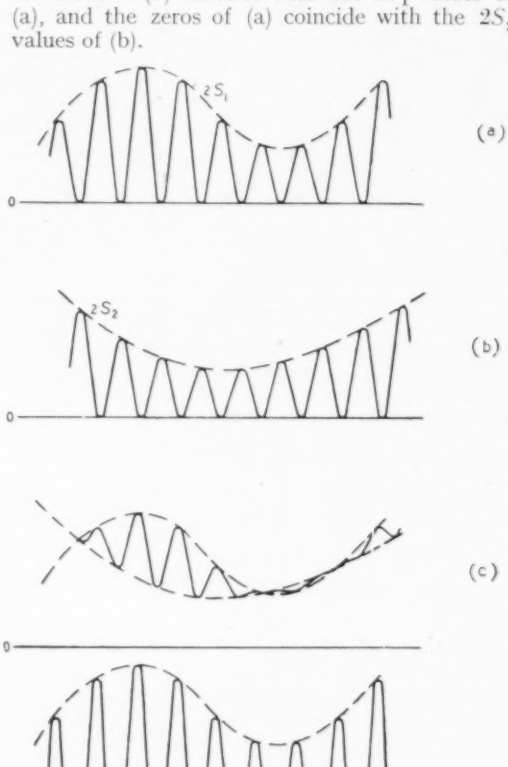


Fig. 23. *(a) The waveform of the signal S_1P_1 after the transmission filter in a dot-interlaced system for two signals; (b) the waveform of the signal S_2P_2 after the transmission filter; (c) the transmission signal obtained by combining (a) and (b); (d) the transmission signal in a dot-interlace system for two signals modified so that the envelopes $2S_1$ and $2S_2$ are separated.*

With two signals in dot-interlace the signal $S_1P_1 + S_2P_2$ is fed to the filter and the transmission signal has the form of Fig. 23(c) if operation takes place according to the principles explained in Section 4. Suppose, however, that instead of feeding the filter with $S_1P_1 + S_2P_2$ it is fed with $S_1P_1 - S_2P_2$. The transmission signal of this modified system is shown in Fig. 23(d); above the zero level the signal contains mainly information about S_1 and below the zero level about S_2.

At the receiver this signal can be separated into two parts by means of a simple clipping device. The outputs have the form shown in Fig. 24 and differ only slightly from the signals represented in Fig. 23(a) and (b). After passing the signals S_1P_1 and S_2P_2 through a filter of bandwidth f_p they are similar to those obtained with synchronous detection, as described in Section 4.2.

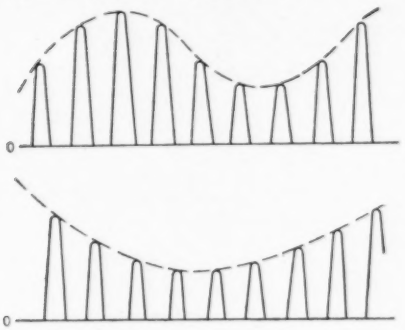


Fig. 24. *The results after clipping the signal of Fig.* 23(*d*).

The signals obtained by this clipping process are not exactly correct, of course, for there will be some cross-talk and distortion. Experiments which the authors have carried out, however, have shown this to be unimportant and matters can be improved by adding correction signals to S_1 and S_2 before modulation.

Comparison of (c) and (d) of Fig. 23 shows that the benefit of non-synchronous operation must be paid for by an increase in the peak value of the cancelling signal in the transmission signal. The peak value in (d) is twice as large as in (c).

Because of the close correspondence between the dot-interlace and sub-carrier systems, it is to be expected that this non-synchronous method of operation could also be employed in the sub-carrier system. In this the upper sub-carrier envelope of the transmission signal is $S_1 + S_2$ and the lower is $S_1 - S_2$, as shown in Fig. 25. The two envelopes, however, are in part located in the same region of signal values, for the envelope of $S_1 + S_2$ varies between 0 and $S_{1max} + S_{2\,max}$ while that of $S_1 - S_2$ varies between S_{1max} and $-S_{2max}$.

Non-synchronous separation, therefore, requires that the transmission signal be modified in such a way that the envelopes are located in different regions of signal values. This can be done by the addition of an amount of sub-carrier, Fig. 26, which must be at least equal to $S_{1max} \cos \omega_p t$, where S_{1max} is the maximum value of S_1.

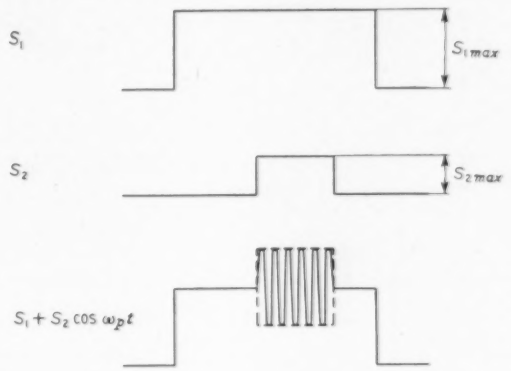


Fig. 25. *The transmission signal in a sub-carrier system for two signals.*

The use of a clipping circuit then produces two signals, Fig. 27, one due to $S_1 + S_2$ and the other to $S_1 - S_2$. As the information of S_1 can be represented by the transmission signal (after suppressing f_p, to decrease the cancelling signal), these signals are needed only to form a signal representing the information of S_2. This in turn can be produced by adding the signals of Fig. 27 and then suppressing f_p. The complete circuit required for all this turns out to be no more than the full-wave rectifier of Fig. 18 and so the use of the clipping method of separation in a sub-carrier system leads to the full-wave rectifying circuit.

7. Survey

The discussion of this article may be summed up as follows:—

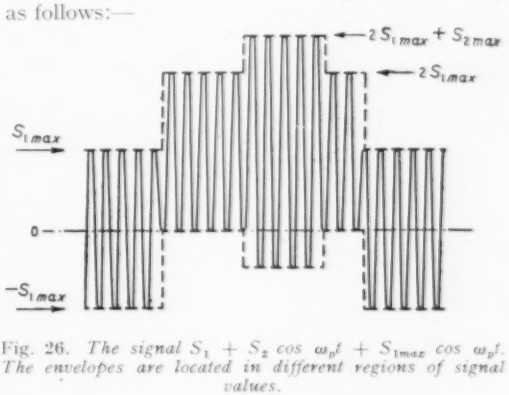


Fig. 26. *The signal* $S_1 + S_2 \cos \omega_p t + S_{1max} \cos \omega_p t$. *The envelopes are located in different regions of signal values.*

If several television signals have to be transmitted simultaneously in the same channel, it can be done by the use of signal components which cancel out in two successive pictures. The original picture can then be reproduced at the receiver by a combination of two successive pictures.

There are two systems based on this principle, the sub-carrier system and the dot-interlace system. They have their own characteristic properties and a preference for one or the other depends on the kind of signals to be transmitted.

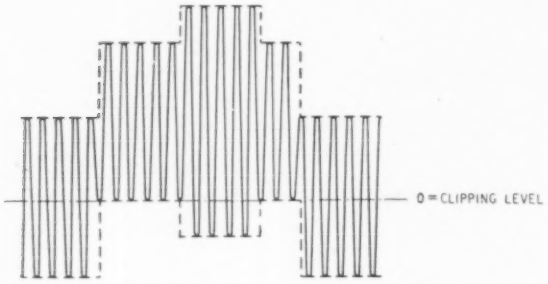


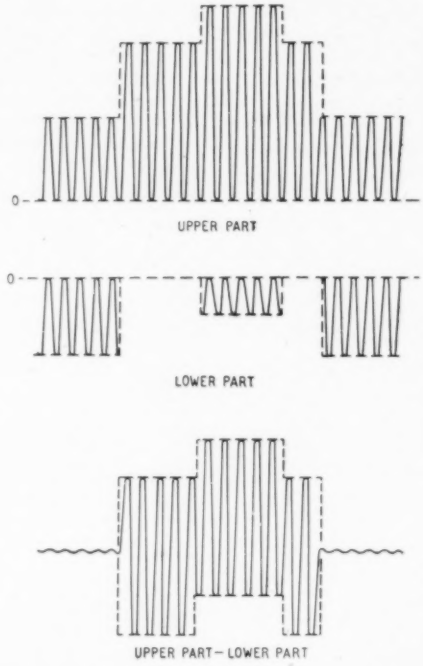
Fig. 27. *The reconstitution of the information of the signal S_2 from the signal $S_1 + S_2 \cos \omega_s t + S_{1max} \cos \omega_p t$ by clipping of the transmission signal and combining the signal parts.*

In the sub-carrier system signals S_1, S_2, S_3, etc., are all transmitted in different ways. S_1 is transmitted normally, but the others are used to modulate sub-carriers of various frequencies, a single sub-carrier being modulated at most by two of these signals. Because of this the sub-carrier system is particularly suited to the transmission of signals of dissimilar qualities; e.g., signals which differ in bandwidth or peak value.

In the dot-interlace system, on the other hand, the signals S_1, S_2, S_3, etc., are all transmitted in the same way. Before transmission they are all transformed by a sampling process which differs for the different signals only in relative phase. The dot-interlace system is, therefore, most suited to the transmission of signals having similar characteristics; e.g., signals which have the same bandwidths and peak values.

With both systems the reconstituted signals at the receiver show an interference effect which cancels out in two successive pictures. In the sub-carrier system this interference with a certain desired signal S_k is related to the undesired signals. As a result, when for some reason cancellation is not fully achieved, there will be cross-talk of the unwanted signal in the desired one. However, when the unwanted signal is zero over some period no interference at all will be experienced.

The sub-carrier system is, therefore, a particularly suitable one when it is desired to keep one of the signals as free from interference as possible. In colour television, for example, compatibility requires that there should be a minimum of interference with the normal transmitted signal, which is often called the luminance signal and corresponds to a black-and-white picture. The ultimate result, too, is not very sensitive to mutual cross-talk because there is a large degree of correlation between the three signals which represent a colour picture.



In the dot-interlace system, however, interference with the separated signals at the receiver is related only to the desired signals themselves. Cross-talk effects are therefore absent, in principle, but the interference effect is always present in the same magnitude in all signals. Because of this the system is most suited to cases where the signals are very sensitive to cross-talk, as when they are entirely uncorrelated.

In some special forms of the sub-carrier system, e.g. if a sub-carrier is modulated by two signals, and in the original form of the dot-interlace system, a synchronous detector must be used at the receiver for separating the signals. This is, in effect, the multiplication of the transmission signal by a waveform of special frequency, phase and shape. By modifying the transmission signal suitably, however, a non-synchronous method of separation can be used with either system.

Non-synchronous separation, however, is restricted to the case where each sub-carrier of a sub-carrier system is modulated by one signal only. In a dot-interlace system it is restricted to the case where two signals only are transmitted.

In general, the use of non-synchronous instead of synchronous separation simplifies the receiver but increases the peak value of the transmission signal, especially the peak value of the cancelling components of this signal.

In the sub-carrier system, however, this increase of the cancelling components can be limited by using a special double-detection circuit, provided that the bandwidths of the signals modulated on the sub-carrier are small compared with the sub-carrier frequency. Under these conditions, the system may be a happy compromise between the disadvantage of increased peak value and the advantage of simple receiver circuitry.

With dot-interlace, separation can be carried out very simply if the modified system for two signals is used, for the cathode-ray tube itself can act as the clipping device. It is possible that in industrial or commercial television the situation may arise where the multiplex transmission of two signals is required, and this simple separation method may well be preferred to compatibility and a small peak value for the transmission signal.

The most important properties and consequences of both the sub-carrier and the dot-interlace systems have now been mentioned. It is obvious that transmission systems could be designed which combine some properties of both. For example, the transmission of two signals of the same bandwidth and a third of smaller bandwidth could be achieved by two dot-interlaced signals with a modulated sub-carrier for the third, the sub-carrier being outside the band occupied by the dot-interlaced signals.

So far the most important application of the multiplex transmission of television signals is to colour television. When compatibility is needed, it appears that the sub-carrier system is the better, because of the following characteristics of colour television:—

(1) The colour picture information is most advantageously transmitted by three signals of which one has the full channel bandwidth and the other two much smaller bandwidths.

(2) Compatibility requires a cancellation component of the transmission signal which has as small a peak value and as high a frequency as possible.

(3) The correlation which exists between the three signals reduces the sensitivity to mutual cross-talk.

It is considered, therefore, that the sub-carrier system should be chosen for a compatible colour television system. It is less clear, however, whether synchronous or non-synchronous operation is to be preferred. This question can probably only be settled after extensive experiments. It should be noted that the use of non-synchronous separation requires two sub-carriers in the transmission signal, whereas with synchronous separation a single sub-carrier modulated by two signals can be employed.

Finally, it should be pointed out that the principle of dot-interlace, as described in this article, can also be used for the transmission of one signal of large bandwidth through several parallel transmission paths of relatively small bandwidths. It could, for instance, be used when it is desired to record a television signal on magnetic tape using several tracks.

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NYQUIST'S CRITERION

A Simple Derivation

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SUMMARY.—Nyquist's form of the stability criterion is deduced using well-known properties of complex numbers and without using integration in the complex plane. A new form of criterion—the 'minimum-phase criterion'—is also developed.

1. Introduction

THE usual proofs[1,2] of Nyquist's stability criterion assume a knowledge of contour integration, and are based upon a theorem in the theory of functions of a complex variable. It is shown in this paper that Nyquist's form of the stability criterion can be deduced, using well-known properties of complex numbers and without using contour integration in the complex plane.

2. Stability of Linear Electrical Networks

The external gain of a single-loop feedback amplifier, G say, is given by

$$G = \frac{A}{1 - A\beta} \quad .. \quad .. \quad (1)$$

where A and β are the gains of the amplifier and feedback portions respectively. If the system be analysed in the steady-state with all radio valves replaced by their equivalent circuits, then A, $A\beta$, $1 - A\beta$ and G can each be expressed as the ratio of two polynomials in p ($p = j\omega$) with real coefficients. Values of p that make the numerator polynomial of a network function vanish, are said to be the zeros of the network function and values of p that make the denominator polynomial vanish, are said to be the poles of the network function. If a network function is of the form response/drive, then the poles of the network function are the natural modes of oscillation of the network. These natural modes of oscillation are sinusoids or exponentially increasing or decreasing sinusoids, if the network is composed of linear elements, and are excited when a drive is applied. Mathematically, the response to a drive is shown by Heaviside's Expansion Theorem or the Laplace transform method to contain the algebraic sum of quantities consisting of a constant with respect to time multiplied by e^{pt} where p takes on values equal to the poles of the network function. The poles of the network function will be complex, in general, and can be plotted as points on the complex p-plane. For poles on the real p-axis, the quantity e^{pt} will be an exponentially increasing or decreasing function of time according to whether p is positive or negative respectively, while for a pair of conjugate complex poles, a pair of the quantities e^{pt} will yield an exponentially increasing or decreasing sinusoid, according to whether the real part of p is positive or negative respectively. Hence, for the network to be stable the natural modes of oscillation must exclude increasing exponentials and exponentially increasing sinusoids, and so the criterion for stability is that no poles of the network gain function, can lie in the right-half p-plane.

3. Stability of Single-Loop Feedback Amplifiers

Consider now the possible positions of the poles of the functions appearing in Equ. (1). As a passive network must be stable, the poles of β must lie in the left half p-plane. The amplifier portion of a single-loop feedback amplifier consists of a series of passive networks connected by radio valves, which serve to separate the passive networks and to drive the input of one of the passive coupling networks with a voltage equal to μ times the output voltage of the previous network. The anode a.c. resistance of the valve is included in the passive network connected to the anode, and μ is the amplification factor of the valve. Thus the poles of the amplifier portion, A, of the feedback amplifier are the total poles of the individual passive networks and must therefore lie in the left-half p-plane. Thus the amplifier portion of a single-loop feedback amplifier is stable on its own. Since, therefore, the poles of G consist of the poles of A and the zeros of $(1 - A\beta)$, it follows that the criterion for stability of a single-loop feedback amplifier is that none of the zeros of $(1 - A\beta)$ should lie in the right-half p-plane.

Further, since poles of $(1 - A\beta)$ and $A\beta$ coincide, for $(1 - A\beta)$ becomes infinite only when $A\beta$ is infinite and vice versa, and neither A nor β has poles in the right-half p-plane, it follows that $(1 - A\beta)$ can have no poles in the right-half p-plane and hence the necessary and sufficient condition for stability of a single-loop amplifier is that $(1 - A\beta)$ (i.e., 1-loop transmission) should be a minimum-phase* function.

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* A network function is said to be minimum-phase if it has no poles or zeros in the right-half p-plane (see ref. 2).

4. Nyquist's Criterion

$(1 - A\beta)$ can be written,

$$(1 - A\beta) = K \frac{(p - p_1)(p - p_2) \ldots (p - p_n)}{(p - p_1')(p - p_2') \ldots (p - p_m')}$$

where K is a real constant; $p_1, p_2, \ldots p_n$ are the zeros and $p_1', p_2', \ldots p_m'$ are the poles of $(1 - A\beta)$. Thus,

phase $(1 - A\beta)$ = phase K + phase $(p - p_1)$ + phase $(p - p_2) + \ldots +$ phase $(p - p_n)$ − phase $(p - p_1')$ − phase $(p - p_2') - \ldots -$ phase $(p - p_m')$.

Consider now the net phase change in $(1 - A\beta)$ as p varies round the path in the complex plane shown in Fig. 1.*

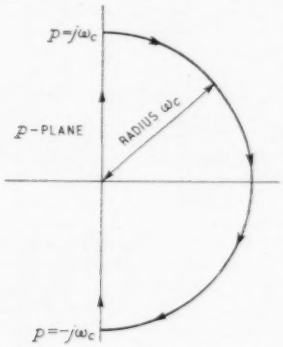

Fig. 1. *Path of variation of p.*

Since phase $(p - p_1)$ is the angle made by the vector joining p_1 to p with the positive real axis, the net phase change for poles and zeros in the left-half p-plane, as p traverses the path shown in Fig. 1, is zero. If the path shown in Fig. 1 is made sufficiently large to include any zeros there may be in the right-half plane and as $(1 - A\beta)$ can have no poles in the right-half p-plane, the net phase change in $(1 - A\beta)$ as p traverses the path shown in Fig. 1 is equal to -2π multiplied by the number of zeros of $(1 - A\beta)$ contained in the right-half p-plane. Thus, the number of times a plot of $(1 - A\beta)$ encircles the origin in a clockwise direction is equal to the number of zeros of $(1 - A\beta)$ in the right-half p-plane. For a practical amplifier circuit $|A\beta|$ will fall to zero, owing to shunt capacitances, as $\omega \to \infty$ so the plot of $(1 - A\beta)$ as p travels round the path in Fig. 1 for $\omega_c \to \infty$ becomes a plot of $(1 - A\beta)$ for real angular frequency variations of from $-\infty$ to $+\infty$ or a plot of $(1 - A\beta)$ and its conjugate for real angular frequencies of from 0 to $+\infty$. For, as the network is composed of real elements, the polynomials of which $(1 - A\beta)$ is composed have real coefficients and so the value of $(1 - A\beta)$ at a given negative real frequency is equal to the conjugate of its value at the corresponding positive real frequency. Thus we have the Nyquist criterion for the stability of single-loop feedback amplifiers:

> If the plot of $(1 - A\beta)$, or $-A\beta$, or $A\beta$ and its complex conjugate for all frequencies from 0 to $+\infty$ encloses the points (0, 0), (−1, 0) or (1, 0) respectively, the resultant single-loop feedback amplifier will be unstable, otherwise not.

5. The Minimum-Phase Criterion

The 'minimum-phase criterion' for the stability of single-loop amplifiers, derived in Section 3, is an alternative formulation to Nyquist's criterion. The minimum-phase criterion might be formulated as follows:

> Plot the modulus of $(1 - A\beta)$ against frequency and obtain the corresponding minimum-phase characteristic by one of the methods given in the literature.† Compare this minimum-phase characteristic with the phase/frequency characteristic of $(1 - A\beta)$—any discrepancy indicates instability of the resultant single-loop amplifier.

Of course, if both the A and β circuits of the amplifier were known to be minimum-phase, no phase measurements would be needed on the apparatus, as with Nyquist's criterion. The shape of the $|1 - A\beta|$/frequency or phase $(1 - A\beta)$/frequency curves is an indication of the nearness of zeros of $(1 - A\beta)$ to the right-half p-plane.

APPENDIX

The following point may be of interest. If the path in Fig. 1 had been the imaginary p-axis and an infinite semi-circle in the left-half p-plane instead of in the right-half p-plane then the same Nyquist diagram would have resulted (for $A\beta \to 0$ as $\omega \to \infty$) regardless, apparently, of the number of zeros of $(1 - A\beta)$ in the right-half p-plane. From the point of view of the infinite left semi-circle, the number of times the Nyquist diagram for $(1 - A\beta)$ encircles the origin in a clockwise sense indicates the number of poles minus the number of zeros of $(1 - A\beta)$ in the left-half p-plane. From the point of view of the infinite right semi-circle, the number of times the Nyquist diagram for $(1 - A\beta)$ encircles the origin in a clockwise sense indicates the number of zeros minus the number of poles of $(1 - A\beta)$ in the right-half p-plane. But for $A\beta \to 0$ as $\omega \to \infty$, $(1 - A\beta)$ has equal numbers of poles and zeros and so the excess of poles over zeros in the left-half p-plane equals the excess of zeros over poles in the right-half p-plane.

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[1] H. Nyquist, "Regeneration Theory", *Bell System Technical Journal*, January 1932, Vol. 11, p. 126.

[2] H. W. Bode, "Network Analysis and Feedback Amplifier Design" (D. Van Nostrand, 1945).

[3] W. T. Duerdoth, "Some Considerations in the Design of Negative-Feedback Amplifiers", *Proc. Instn elect. Engrs*, Part III, 1950, Vol. 97, pp. 138–155.

[4] O. P. D. Cutteridge, "A Graphical Contribution to the Analysis and Synthesis of Electrical Networks", 1953, I.E.E. Monograph No. 64.

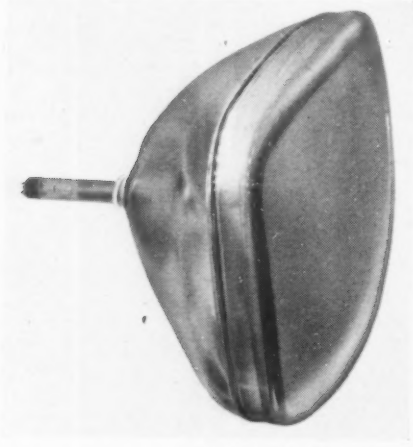
* A semi-circle in the left half-plane could also be used—see Appendix.

† See, for example, refs. 2, 3 or 4.

21ST NATIONAL RADIO EXHIBITION

THIS year's exhibition was again held at Earls Court and was open from 24th August to 4th September. As usual, it was predominantly one of broadcast receiving equipment, including television. In the main hall the exhibits were almost entirely in the domestic category, but in the gallery there were large displays by the three Services, a small one of electronic apparatus and one devoted to aspects of technical training. Here, too, the enormous B.B.C. show was staged.

Since the exhibition is designed mainly to attract the non-technical public it is, perhaps, natural that the emphasis should have been on appearance, price and performance. But it is curious to observe that, in spite of the efforts of standardization committees and the like, television has now no connection with radio! To the public, radio means sound broadcasting; television is something quite different! There is room for a great deal of misunderstanding here, for the engineer, in referring to radio and correctly including within it television as a special branch, will now certainly be misunderstood.



Ediswan 27-in c.r. tube, type CRM271.

It is by common usage that a language develops and it is probably impossible to fight against this trend, much as engineers may dislike it. It looks very much as if a new word will be wanted soon for the present correct meaning of radio; or will it be a return to 'wireless' as the general term?

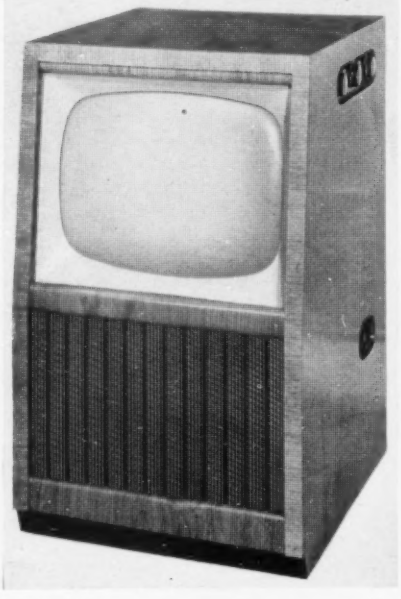
Both from the engineering and the general public's points of view, this year's exhibition is noteworthy for two things which mark a breakaway from the trends of the past. One is in sound broadcast receivers, the other is in television. Both are due to the decisions to provide transmissions on new frequencies. In the one case it is the decision to supplement the existing sound-broadcasting transmissions by f.m. stations on Band II (82—100 Mc/s); in the other it is the decision to provide alternative television programmes in Band III (174—216 Mc/s).

This is not, however, the only trend in television. The most cursory inspection of the exhibits made it very clear that the general trend is towards larger pictures. The use of 14–17-in tubes is now usual and the emphasis seems to be on the 17-in rectangular type. The result was to make the relatively few sets with 12-in tubes look very small; one was inclined to mistake them for 9-in but, actually, there were no 9-in shown! The trend was made more noticeable by a number of even larger tubes which were shown by the tube makers or demonstrated in prototype apparatus. Of these, the Ediswan CRM 271 is noteworthy for a screen diagonal of 27 in.

Above 17-in, the tendency is to increase the deflection angle to as much as 90° so that the total length of a 21-in tube is no more than that of the ordinary 17-in. This entails very careful deflector-coil design if a uniform focus is to be secured and it appears to be necessary to design the coil for the best focus irrespective of what happens to the raster shape. The result is usually a pincushion raster which must be corrected by other means, as by small magnets around the face of the tube.

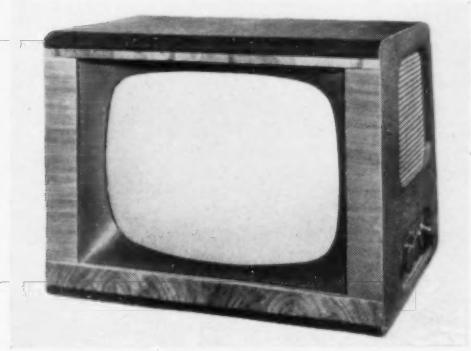
In their demonstration room, Mullard were showing a 21-in tube with a deflection-assembly of the so-called castellated-yoke type but with 16 slots instead of the usual 8. This enables a two-step graded winding to be used in place of the usual lumped winding, and so a more uniform field distribution can be obtained.

Among valves for television sets, the only major development has been in types for Band-III operation and most makers now have more-or-less equivalent



Bush T36; the controls are in two groups on the right-hand side of the cabinet.

types. A double-troide intended for use in the cascode circuit and a triode-pentode for frequency-changing form the new valves. In the Mullard range there are the PCC84 and PCF80 and similar types in other ranges are the B319 and LZ319 (Marconi and Osram) and 30L1 and 30C1 (Mazda). The valves are of the miniature all-glass type with the B9A base and are specially designed for operation at around 200 Mc/s. They have 0.3-A heaters designed for series operation and the triode has an input damping resistance of 2 kΩ at 200 Mc/s.

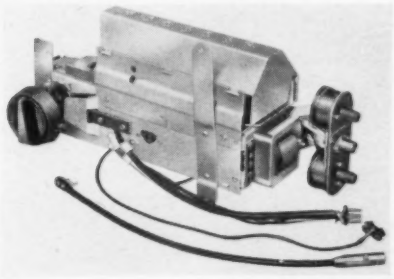


G.E.C. 17-in table model, type BT2745. The loudspeaker grille is on the right of the cabinet.

The need for these valves has only just arisen because hitherto British television has been entirely in Band I (41–68 Mc/s) where the more ordinary types are satisfactory. The advent of Band III has done more than call for new valve types; it has changed the appearance of television sets slightly. Most have grown an extra knob or two for, with alternative programmes looming up, they must now include some kind of station selector. This is a new thing in British television. There have, of course, been five television channels in Band I for a long time but no one has wanted to be able to change rapidly from one to another, since they serve different parts of the country with the same programme. All that has been necessary hitherto is that a set should be capable of being adjusted to receive a different channel in case an owner moves from one part of the country to another. Most set makers went further, it is true, but mainly to avoid having different models for different channels.

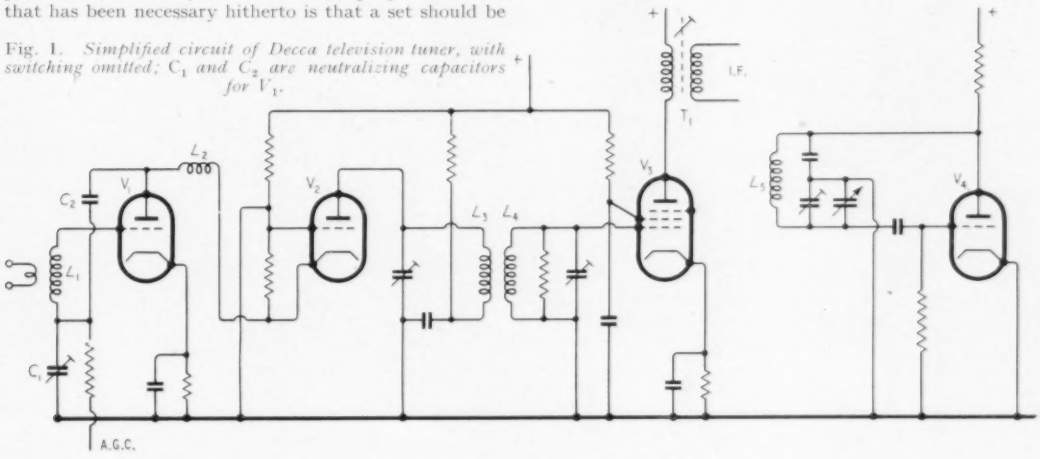
The need for providing a user control to change over from one channel to another is new. Also, there will be five channels in Band I and eight in Band III. The most general solution to the problem of station selection is to provide continuous tuning in both bands with a changeover switch between bands. This is what Bush do. Ganged cores for the coils are operated by two separate knobs, one for each band. Since it is unlikely that there will be more than one station in each band receivable at a given point for a good time to come, one can obviously treat the tuning knobs rather as pre-set controls and use the band switch as an alternative programme selector.

Another system which has some resemblance to this one is used by G.E.C. There are three sets of coils tuned by slugs; one gives continuous coverage of Band I and the other two give continuous coverage of Band III. The slugs are not ganged and are internal pre-set adjustments by which the set can be tuned to any channel in Band I and any two in Band III. A three-position switch enables the user to select any of the three pre-set stations.



Ultra television tuner with three-position selector switch giving one Band I and two Band III channels.

Fig. 1. *Simplified circuit of Decca television tuner, with switching omitted; C_1 and C_2 are neutralizing capacitors for V_1.*



Most set makers, however, have adopted some form of turret tuner having 12 sets of coils for five channels in Band I and seven in Band III. All the coils are not always fitted initially; sometimes there are only three sets and the turret is arranged to act as a three-position switch. In any case, in the true turret, the coil assemblies are units held in by clips and are readily changed.

It is a feature of the turret tuner that the coils for each channel are completely independent of all others. In an alternative in which the turret is replaced by wafer switches the coils are all connected in series and operation of the switch adds incremental inductance. All trimming adjustments must then be done in order, starting with the highest frequency channel. An arrangement of this nature is employed by Pye[1] and on Band III it can be pre-set for channels 6–12 or 7–13 by adjusting a trimmer on the oscillator circuit.

However the coil switching is arranged, the circuit employed is much the same. It is usual to employ a double-triode as a cascode r.f. amplifier with a triode-pentode as a frequency-changer. A typical example is shown in Fig. 1, being a somewhat simplified version of the one used in Decca receivers. The main circuit difference between different tuners lies in the inter-stage coupling; some have a coupled pair of tuned circuits like L_3, L_4 of Fig. 1, others have only one tuned circuit here.

The signal coils L_1, L_3 and L_4 and the oscillator coil L_5 are all switched or changed by a turret. The coil L_2 is not. This is tuned at the h.f. end of Band III and acts to maintain the gain at the highest frequencies. The capacitance C_1 acts with C_2 to neutralize V_1. This is desirable to increase the input impedance of V_1 and it helps to reduce oscillator radiation.

English Electric T40 with 17-in tube. Concentric controls are used.

The output at intermediate frequency is taken via a screened cable of low impedance fed by T_1 so that the tuner can be placed in any convenient position and does not have to be near the i.f. amplifier. The oscillator V_4 is usually a simple Colpitts type and a trimmer is usually brought out as a panel control to permit the user to correct for any oscillator drift.

There has been a tendency for some years towards the use of higher intermediate frequencies, of the order of 34 Mc/s, and the advent of Band III has given this an impetus. Even this frequency is rather low for Band-III requirements but it cannot be higher because of Band I. There is no uniformity, as yet, and quite a lot of different frequencies within one or two megacycles of 34 Mc/s are used. Some people, too, still stick to lower frequencies, 16 Mc/s or so, but there is no doubt that the trend is towards 34 Mc/s or thereabouts.

In the i.f. amplifier itself the use of coupled pairs of tuned circuits as intervalve couplings is on the increase, probably because of the need for better adjacent-channel selectivity. Rejector circuits are invariably included to improve the sound-channel rejection. Three stages in the vision channel are usual, although some sets manage quite well with only two. With the narrower bandwidth of the sound channel no more than two stages is ever needed here.

One rather unexpected result of having to provide for Band-III reception is the inclusion of some form of a.g.c. in many sets. Hitherto, automatic gain control has hardly been necessary in British television and it has not often been used except in models intended for fringe-area reception, where fading is likely. When alternative programmes become available, however, it is hardly to be expected that the two signals will be of the same strength and a.g.c. becomes desirable to take up the variations.

A true a.g.c. system which is strictly analogous to the ordinary kind used in sound receivers is rather difficult in television, for there is no constant mean level to the vision signal. The mean level varies both with the signal strength and the modulation. The only thing that does not vary with the picture content is the black level. A true signal-level a.g.c. system must be operated by the black level, therefore, and this requires the use of an electronic switch arranged so that the signal is connected to the a.g.c. system only when the signal corresponds to black level.

There are a great many different arrangements possible which can be divided broadly into two classes, those gated at line frequency and those gated at frame frequency. There is an interval of some 6 μsec of black level, known as the back porch, after each line sync pulse and the signal can be gated during this interval by a pulse derived from the line time-base. Alternatively, the gating can be at frame frequency by a pulse derived from the frame time-base for, during the frame flyback period, the signal is in the form of pulses varying between zero and black level.

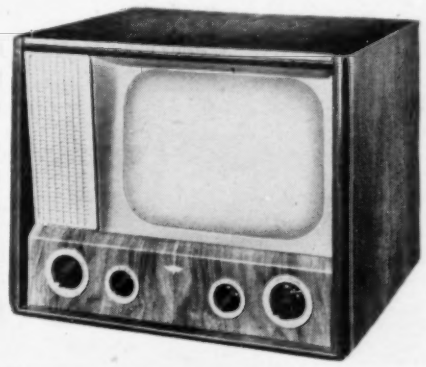
Ultra have adopted the frame-gating system, while Murphy now use the line-gating method, as do Pye. The circuits employed differ considerably, however, but in all cases two or three extra valves are needed.

Gated a.g.c. systems enable a fading signal to be dealt with as well as the variations between alternative signals without black level being affected. When only alternative programmes need be considered one very simple arrangement is possible, which is to control the video channel gain from the a.g.c. system of the sound channel. It depends for its success upon the ratio of the sound to the vision signals being the same for the two

channels and this appears to be a reasonable assumption within the service area. English Electric use this arrangement.

Most set makers, however, use a fairly simple form of a.g.c. operating from the mean level of the video signal. This does, of course, tend to keep the mean brightness of the picture constant and so it is theoretically a disadvantageous arrangement. Its advocates, however, argue that the mean brightness of a television picture does not, in fact, vary very much and the loss of artistic effect on the rare occasions when it does is outweighed by the great simplicity of the method.

In every set, a suitable signal is developed on the grid of the sync separator, for the d.c. restoring action makes the mean grid potential proportional to the mean signal level. Basically, therefore, it is necessary only to filter out the signal variations by an RC network to have an a.g.c. supply. In some cases, Bush for instance, this is all that is done. In others, a diode is added to clamp the minimum bias voltage at a definite value. In still others, although the principle is the same, the sync separator grid voltage is not used but a separate diode is employed to develop the a.g.c. voltage; H.M.V. adopt this arrangement in the 1828 fringe-area model.



Kolster-Brandes LVT30 with 14-in tube. The controls on the left are station-selector and fine-tuning.

The tendency of these circuits to maintain a constant mean brightness of picture means that the black level is no longer fixed. In order to prevent the frame flyback lines from showing on certain pictures, therefore, a pulse is applied to the c.r. tube to black them out. This is usually a negative-going pulse from the frame time-base which is applied to the grid of the tube; sometimes, however, it is a positive pulse applied to the cathode. A simple RC shaping network is often included between the two.

The use of this flyback suppression circuit is by no means confined to sets which are fitted with a.g.c. Its use is becoming quite general in all sets. There is no doubt that it is desirable unless the black level is maintained perfectly and the desirability of doing this seems a very debatable matter. Many, if not most, designers introduce some deliberate attenuation of the d.c. component of the signal.

Flywheel sync, which is universal in the U.S.A., is very little used in British practice, probably because the different characteristics of positive modulation render it unnecessary in areas of fairly high field strength. The line time-base is locked directly from the differentiated sync pulse in most receivers. However, in fringe areas, signals are weak and noise and interference are much more serious, and then flywheel sync is undoubtedly advantageous. Consequently, it has become a usual fitting in fringe-area models. The actual circuits used vary considerably in the different sets, however.

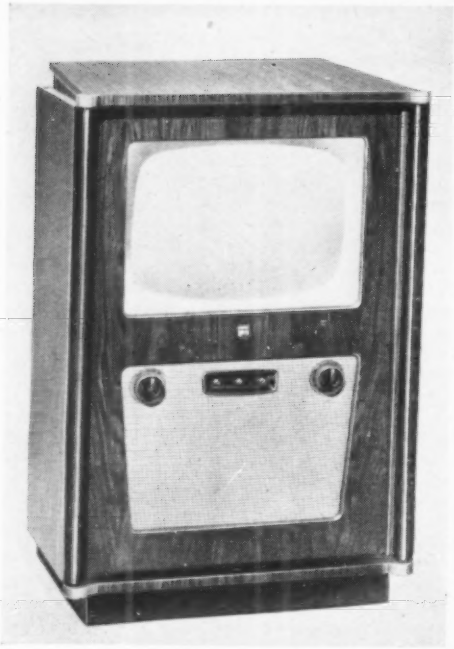
Little change has occurred in time-base practice during the year. There is a somewhat increased tendency towards the use of negative feedback for linearizing the frame time-base, and towards the adoption of a partially-saturated reactor in series with the deflector coils for linearizing the line scan. Flyback e.h.t. is invariably employed in sets for direct viewing but in projection sets an a.f. oscillator is used. Projection has made little or no headway in the domestic type of equipment and there are relatively few sets of this type. Ferranti make one giving a picture of 20-in diagonal which is unusual in having a push-pull video stage to drive the c.r. tube. One valve feeds the signal to the grid and the other to the cathode of the tube.

In general, however, projection now finds its main application in providing pictures of some 5-ft diagonal. Front projection is usually adopted, as in the Nera sets and in the White-Ibbotson.

Most people employ the Mullard $2\frac{1}{2}$-in tube operating at 25 kV with a Schmidt optical system. H.M.V., however, have produced a projection model this year, in which a tube of their own is used; this is a $3\frac{1}{2}$-in type and operates at 28 kV. The viewing screen is unusual in being of aluminium and being curved along its longer dimension so that it is concave towards the viewer. It is claimed that this is done to secure a more uniform focus.

Generally speaking, the permanent magnet has quite displaced the electromagnet for focusing the c.r. beam. There are signs, however, that in time it may in its turn be displaced by electrostatic focusing. This is by no

Chassis of Ambassador TV14/15 receiver. The tuner is fitted at the front of the picture. Pre-set controls are visible at the front under the tube.



Ekco TC209 receiver with 17-in tube. An f.m. receiver for Band II is included.

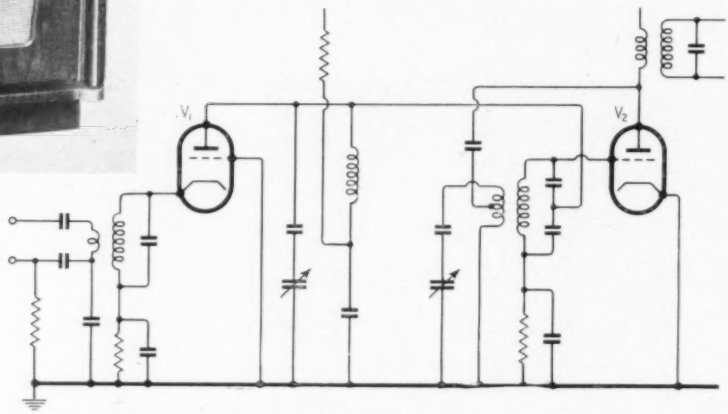


Fig. 2 (right). *Basic circuit of the Kolster-Brandes Band II tuner using one double-triode.*

means a new thing, for H.M.V. had sets embodying magnetic deflection and electrostatic focusing as long ago as 1938.[2]

This year, it is Kolster-Brandes who have adopted it in the LVT 50. The tube used is the Brimar C17JM, a 17-in tube with ion trap and aluminized screen. It operates normally with 14 kV applied to the second and final anodes and 300 V on the first anode. The third anode is the focus electrode and, according to the tube, ratings can be at from −64 V to + 350 V. In the LVT 50 receiver it is returned to chassis and is thus negative to cathode. The set is unique in having no focus control.

Quite a large number of firms exhibited the chassis of their television sets this year and it was quite noticeable how standards of construction have improved. It is obvious that the mechanical design is now given as much attention as the electrical and that manufacturers now recognize that sets do occasionally go wrong! Not everyone deals with the maintenance problem, which is largely one of accessibility, in the same way, of course, but it is obvious that it has received serious consideration.

In some cases, almost everything is assembled on a single chassis of conventional form, but the parts are so cunningly disposed that there seems to be nothing that cannot be reached at once. Others obtain accessibility by adopting a unit construction, it being intended that a unit should be removed if it becomes necessary to delve inside it. The relative merits of the two systems are debatable, but there seems little doubt that, however it has been achieved, this year's television sets will be easier to maintain than the older ones.

This year there are larger pictures, a better mechanical design and improved accessibility for maintenance. There is a tendency, too, although less general, to fit the subsidiary controls at the front beneath a concealing flap instead of at the back. Everyone will welcome this change. When they are at the back, it is all too often quite impossible even to touch a control, much less operate it, and at the same time to see the picture! The front position for the controls benefits not only the user but also the dealer who installs the set, and is likely to result in better pictures since it becomes easy to adjust the controls.

In the sphere of sound broadcasting, major develop-

An unusual form of construction is adopted in the McMichael Clubman radio-gramophone.

ment has occurred mainly through the need for providing for f.m. reception on Band II. The usual practice seems to be to employ a double triode in a circuit of the form shown in Fig. 2. One section operates as an earthed-grid amplifier and the other as a mixer-oscillator. The output i.f. is at 10·7 Mc/s and is fed to the heptode section of the medium and long waveband mixer which is turned into an extra i.f. stage for v.h.f. reception. For normal broadcasting the sets have a frequency-changer, one i.f. stage at 465 kc/s or thereabouts, a diode detector and a two-stage a.f. amplifier. On v.h.f. the mixer becomes an extra i.f. stage, the frequency is changed to 10·7 Mc/s, and the detector circuit is altered to that of a ratio detector. No special limiter is thus needed.

That is the general pattern of the Band-II receiver as shown this year. There were not, however, a great many on view so the pattern may not be representative of the industry when the reality of Band II gets a little nearer. An exception to the general run is the Dynatron v.h.f. set, for this is pre-tuned with switch station selection. A limiter is used with a Foster-Seeley discriminator and a.f.c. is employed to keep the set in tune.

One thing that follows inevitably from the adoption of Bands II and III is the need for new aerials. For Band II the requirements are straightforward and an aerial is much the same as an ordinary Band I television aerial scaled down for the new frequency and turned on its side to suit horizontal polarization.

Band III problems would be equally straightforward were it not for the fact that the aerial must be combined with one for Band I. Very few such combination aerials were shown, for the fact is that it is too early to start designing them. Since the location and power of the Band III stations has not yet been decided, it is not yet known what aerial gain will be needed nor whether the beams on the two bands will be in the same direction or will have to be adjustable to lie at any angle.

One thing is certain; feeder losses will be higher on Band III than on Band I. With this in mind, some set makers provide for the use of a 300-Ω twin feeder for Band III while retaining the usual 72-Ω coaxial cable for Band I. However, a new coaxial cable shown by Aerialite is of particular interest in this connection. It is of the usual general form with polyethylene insulation but has five longitudinal holes in the insulant parallel with the wire to give a partial air-dielectric. Known as the Aeraxial cable, it is available in two forms, one having 2·1-db attenuation per 100 ft at 45 Mc/s and 4·5 db at 200 Mc/s, and the other 1·2 db and 2·6 db respectively, with an overall diameter of under $\frac{1}{4}$ in.

The complexity of modern radio apparatus calls for the use of quite elaborate equipment for maintenance purposes. A certain amount of test gear was shown but not a great deal, quite naturally, since this exhibition is mainly one for the public. What was shown, therefore, is in no way representative of this branch of the industry. Taylor showed a wobbulator (TV Sweep Oscillator, Model 92A) for use with an oscilloscope; it has a mean frequency adjustable up to 250 Mc/s so that it can be used at any signal frequency in Bands I and III or to feed into any i.f. amplifier.

Taylor Electrical television wobbulator.

Waveforms showed pattern generators for television set testing, the Radar 405, which covers both television bands and provides the correct synchronizing waveform together with a selection of test patterns. The Cossor model 1322 is a combined wobbulator and marker generator. It has a 10-Mc/s sweep with a mean value adjustable from 5 Mc/s to 225 Mc/s in four bands and there is a 2-Mc/s crystal calibrator included.

REFERENCES

[1] "12-Channel Television Tuner", *Wireless World*, April 1954, p. 162.
[2] "The Wireless Exhibition, 1938", *Wireless Engineer*, October 1938, p. 543.

POST OFFICE APPOINTMENTS

Sir Gordon Radley, C.B.E., Ph.D.(Eng.), who was Engineer-in-Chief, G.P.O., has been appointed Deputy Director-General. To deal with broadcasting, an additional and temporary post of Deputy Director-General has been created and will be filled by Sir Ben Barnett, K.B.E., C.B., M.C. Brigadier L. H. Harris, C.B.E., Controller of Research, succeeds Sir Gordon Radley as Engineer-in-Chief, and G. J. S. Little, C.B.E., G.M., an Assistant Engineer-in-Chief, becomes Controller of Research.

BRIT.I.R.E.

The Council of the British Institution of Radio Engineers has announced that the next President is to be Rear-Admiral Sir Philip Clarke, K.B.E., C.B., D.S.O. He succeeds W. E. Miller, M.A.(Cantab.), who has been President for two years in succession.

The Annual General Meeting will be held at London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1, on 27th October at 6 p.m., followed by the Presidential Address at 7 o'clock.

I.E.E. MEETINGS

7th October. President's Inaugural Address by J. Eccles, C.B.E., B.Sc.

12th October. "The Teaching of the Subject of Electromagnetism", discussion to be opened by K. J. R. Wilkinson, D.Sc., at 6 o'clock.

13th October. Address by the Chairman of the Radio Section, C. W. Oatley, M.A., M.Sc.

25th October. "Whether Compatibility is necessary for a Colour Television System in Great Britain", discussion to be opened by E. P. Wethey.

4th November. "A Transatlantic Telephone Cable", by M. J. Kelly, Ph.D., Sir Gordon Radley, C.B.E., Ph.D.(Eng.), G. W. Gilman and R. J. Halsey, B.Sc.(Eng.).

These meetings will be held at Savoy Place, London, W.C.2, at 5.30 except where otherwise stated.

CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

External Field of a Widely-Wound Toroid

SIR,—In the second Editorial of the June 1954 issue of *Wireless Engineer* you discussed this interesting problem and gave what seems to be on first sight an attractive graphical analysis. I am afraid, however, that this explanation has some of the alluring qualities of a will o' the wisp and may lead your readers into dangerously marshy ground. May I therefore put forward a more detailed analysis, which I hope will be of interest to your readers?

In approaching the problem of the toroid it is first helpful to consider the magnetic field of a long current filament (Fig. 1) which is placed at a distance R from the origin of co-ordinates. This magnetic field can be expressed in terms of a scalar potential function V, which will have to satisfy Laplace's equation. In cylindrical co-ordinates this equation is

$$\frac{\partial^2 V}{\partial r^2} + \frac{1}{r}\frac{\partial V}{\partial r} + \frac{1}{r^2}\frac{\partial^2 V}{\partial \theta^2} = 0 \quad .. \quad .. \quad .. \quad (1)$$

A convenient form of the solution of this equation for $r > R$ is:

$$\frac{V}{2I} = \theta + \sum_{-\infty}^{+\infty} A_n r^n \sin n\theta + \sum_{-\infty}^{+\infty} B_n r^n \cos n\theta \quad (2)$$

In order to satisfy the physical condition that the magnetic force, and therefore the potential, must decrease with distance from the filament, the constant n in the solution must be negative. In order that the magnetic force may be single valued n must be integral and by symmetry cosine terms must be excluded from the solution for V.

We can thus write:

$$\frac{V}{2I} = \theta + \sum_{1}^{\infty} A_n \frac{\sin n\theta}{r^n} \quad .. \quad .. \quad .. \quad (3)$$

The coefficient A_n can be obtained by considering the magnetic force at a particular point. Thus in Fig. 2 the magnetic force H_θ at P $(r, 0)$ is given by

$$-\frac{H_\theta}{2I} = \frac{1}{r-R} = \frac{1}{r}\left(1 - \frac{R}{r}\right)^{-1}$$
$$= \frac{1}{r}\left[1 + \frac{R}{r} + \frac{R^2}{r^2} + \ldots\ldots\right] \quad .. \quad .. \quad (4)$$

But $$-\frac{H_\theta}{2I} = \frac{1}{r}\frac{\partial V}{\partial \theta}\frac{1}{2I} \quad .. \quad .. \quad .. \quad .. \quad (5)$$

Hence

$$\left[1 + \frac{R}{r} + \frac{R^2}{r^2} + \ldots \frac{R^n}{r^n} + \ldots\right]$$
$$= \left[1 + \frac{A_1}{r} + \frac{2A_2}{r^2} + \ldots \frac{nA_n}{r^n} + \ldots\right] \quad .. (6)$$
$$\therefore A_n = \frac{R^n}{n}$$

and the solution of Laplace's equation for our problem becomes

$$\frac{V}{2I} = \theta + \sum_{1}^{\infty} \frac{1}{n}\left(\frac{R}{r}\right)^n \sin n\theta \quad .. \quad .. \quad (7)$$

We have thus obtained an expression for the potential at points r, θ of a current filament distant R from the origin, it being understood that $r > R$.

Consider now the work done in carrying a unit pole around any path enclosing the filament. Such a path can be split up into a number of elemental paths for which either θ or r is constant. Reference to Fig. 3 will show that the work done in moving from A to B is the same as is done in moving from A to C to D to B (since ABDC does not enclose a current). Hence it can be seen that the work done in moving around any closed path is equal to the work done in moving around the circle of radius r $(r > R)$. This work is

$$\int_0^{2\pi} -H_\theta\, r\, d\theta = \int_0^{2\pi} \frac{\partial V}{\partial \theta}\, d\theta = \Big[V\Big]_0^{2\pi} = 4\pi I \quad .. \quad (8)$$

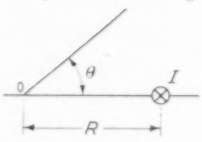


Fig. 1

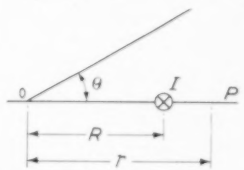


Fig. 2

This result is of course well known, but it is not always appreciated how it comes about that the 'work law' is independent of the path. In the case that we have considered the field is very far from uniform around the path chosen. The strength of this field at any point can be calculated by the use of equation (1). It is at first sight surprising that the line integral should be independent of the path, when the field strength at a point depends so much on the values of the co-ordinates at that point. The mathematical 'mechanism' by which this surprising result comes about can be seen by reference to equation (7), for the field strength is given by an expression involving a θ term and a Fourier series in $\sin n\theta$. All the terms of the Fourier series must disappear when an integral from 0 to 2π is taken and we are thus left with the θ term, which describes the field of a current placed at the origin of co-ordinates. Hence the work done around any path takes no account of the displacement of the filament from the origin but is the same as if an identical current filament were situated at the origin. Of course we should have obtained an identical result by considering the flux associated with a unit pole, since this flux would have to be cut when the pole is carried around the conductor. But the introduction of the concept of flux merely deepens the mystery and tells us nothing of the force distribution around the filament.

Consider now the external field of a number of current filaments disposed at equal intervals around the circumference of a circle of radius R. Let there be p filaments, then the potential function for a point defined by the polar co-ordinates r, θ $(r > R)$ is given by

$$\frac{V}{2I} = p\theta - (p-1)\pi + \sum_{k=1}^{\infty} \frac{1}{k}\left(\frac{R}{r}\right)^{kp} \sin kp\,\theta \quad .. \quad (9)$$

If the total current $pI = I'$

$$\frac{V}{2I'} = \theta - \left(\frac{p-1}{p}\right)\pi + \sum_{k=1}^{\infty} \frac{1}{kp}\left(\frac{R}{r}\right)^{kp} \sin kp\,\theta \quad (10)$$

and $$-\frac{H_\theta}{2I'} = \frac{1}{r}\frac{\partial V}{\partial \theta}\frac{1}{2I'} = \frac{1}{r} + \sum_{k=1}^{\infty} \frac{1}{r}\left(\frac{R}{r}\right)^{kp} \cos kp\,\theta \quad (11)$$

Hence as $p \to \infty$, $$-\frac{H_\theta}{2I'} \to \frac{1}{r} \quad \ldots \quad \ldots \quad \ldots \quad (12)$$

and this is the correct result for a tube carrying a uniform steady current.

As was to be expected, the addition of the fields of a number of current filaments produces a tangential magnetic force that is more nearly uniform (for any given radius) than is the tangential magnetic force of

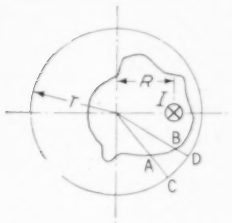


Fig. 3

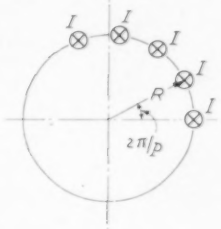


Fig. 4

a single filament. In the limiting case of a tubular current this field is absolutely constant and has the same value as the field of a current of equal magnitude concentrated at the origin of co-ordinates.

The radial force is given by:

$$\frac{H_r}{2I'} = -\frac{\partial V}{\partial r}\frac{1}{2I'} = \sum_{k=1}^{\infty}\frac{1}{r}\left(\frac{R}{r}\right)^{kp} \sin kp\,\theta \quad \ldots \quad (13)$$

and this force disappears when the filaments coalesce to form a tube of current ($p \to \infty$).

Consider now the numerical example illustrated by Fig. 5. It is required to find the tangential magnetic force at a radius given by $r = 1{\cdot}1\,R$. Employing our previous notation we have $p = 8$, and H_θ can be found by the use of equation (11).

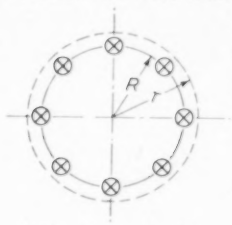


Fig. 5

At $\theta = 0$

$$-\frac{H_\theta}{2I'} = \frac{1}{r}\left[1 + \frac{1}{(1{\cdot}1)^8} + \frac{1}{(1{\cdot}1)^{16}} + \frac{1}{(1{\cdot}1)^{24}} + \ldots\ldots\right]$$

$$\therefore \quad -\frac{H_\theta}{2I'} = \frac{1}{r}[1 + 0{\cdot}466 + 0{\cdot}217 + 0{\cdot}101 + 0{\cdot}047 + 0{\cdot}022 + 0{\cdot}01 + 0{\cdot}005 + \ldots\ldots]$$

$$\therefore \quad -\frac{H_\theta}{2I'} = \frac{1{\cdot}87}{r}$$

and this will also be the value of H_θ at angles given by

$$\theta = \frac{\pi}{4}, \frac{\pi}{2}, \ldots\ldots \frac{7\pi}{4}.$$

Half-way between the filaments (i.e., at angles given by $\theta = \frac{\pi}{8}, \frac{3\pi}{8}, \ldots\ldots \frac{15\pi}{8}$) we have

$$-\frac{H_\theta}{2I'} = \frac{1}{r}[1 - 0{\cdot}466 + 0{\cdot}217 - 0{\cdot}101 + 0{\cdot}047 - 0{\cdot}022 + 0{\cdot}01 - 0{\cdot}005 + \ldots\ldots]$$

$$\therefore \quad -\frac{H_\theta}{2I'} = \frac{0{\cdot}72}{r}$$

Fig. 6 shows the complete curve of H_θ for this ring of eight filaments plotted against θ, and on the same figure a curve of H_θ against θ has been plotted for a ring of four filaments. This figure shows very clearly that the increase in the number of filaments has increased the order of the harmonics in the curve of H_θ (in our example the order of the harmonics has been doubled) and at the same time the amplitude of the harmonics has been considerably reduced. Thus the tangential magnetic force of a ring of only eight filaments is already tending to that of a tube carrying a uniformly distributed current. In the case of a toroid the spacing is likely to be much closer and is therefore not likely to differ substantially from the field of a tube.

In the Editorial, to which we are referring, it is suggested that the field close to one of the filaments should depend dominantly on that filament. In our case $r = 1{\cdot}1\,R$ and the field due to a single filament at this distance from the origin and on a radius vector passing through the filament would be

$$-\frac{H_\theta}{2I'8} = \frac{1}{r - R} \qquad \therefore \quad -\frac{H_\theta}{2I'} = \frac{11}{8r} \approx \frac{1{\cdot}4}{r}$$

But Fig. 5 shows that $H_\theta \approx 1{\cdot}9/r$ in the case of the eight-filament ring and $H_\theta \approx 3{\cdot}1/r$ in the case of the four-filament ring.

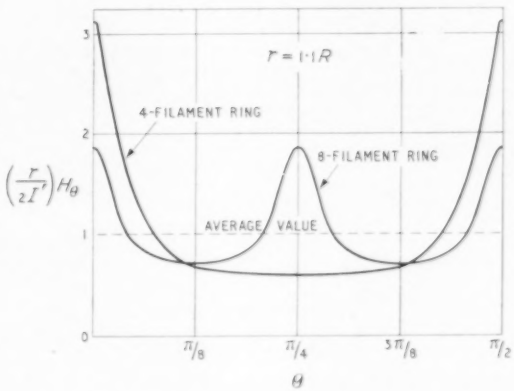


Fig. 6

In calculating H_θ it is thus essential to consider not only the adjacent filament but the complete current distribution. In the interesting case of a cylindrical tube carrying a uniformly distributed current the field is given by $-H_\theta/2I' = 1/r$ and this is exactly twice the magnetic force that would be experienced by a unit pole close to an infinite plane sheet of the same current density as the tube.

Fig. 7 has been derived to show the change in H_θ when the distance r is increased. H_θ has been plotted for a ring of eight current filaments at distances $r = 1{\cdot}1\,R$ and $r = 1{\cdot}2\,R$. It will be seen that the increase in r has the effect of suppressing almost completely all but the first term of the Fourier series for H_θ and the amplitude of this harmonic is also considerably reduced. Hence a ring of current filaments becomes rapidly indistinguishable magnetically from a current-carrying tube, as the distance of the observer is increased.

We have now, in effect, solved the problem of the toroid—at least for the two-dimensional case—because it only remains to superpose the fields of two rings of filaments. It is clear that the external field will have zero average value, if the current flowing in one set of

filaments is equal and opposite to the current flowing in the other set. The analysis has shown that it is not strictly correct to state that there is no external field, but it is true that there is no average external field around any path that embraces both the 'go' and the 'return' current. So we have arrived back at the 'work law'.

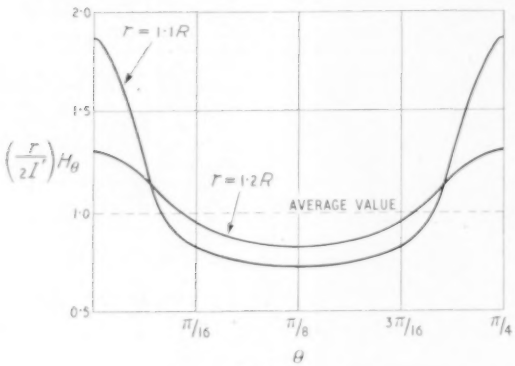


Fig. 7

It should, perhaps, be stated explicitly that our discussion refers only to steady currents, although it will be substantially correct for alternating currents at power frequencies. The complete a.c. solution can readily be worked out, but it is outside the scope of this letter. Reference should be made to a paper entitled "Screening Properties of a Squirrel Cage of Wire", by Professor E. B. Moullin (*J. Instn elect. Engrs*, Vol. 91, Part III, March 1944).

P. HAMMOND.

Department of Engineering,
University of Cambridge.
14th July 1954.

NEW BOOKS

A Text-Book of Radar. (2nd Edition)

By the Staff of the Radiophysics Laboratory, C.S.I.R.O., Australia. Edited by E. G. BOWEN. Pp. 617 + xiii. Cambridge University Press, 200 Euston Road, London, N.W.1. Price 45s.

In this revised edition of a book which first appeared in 1948, only minor alterations have been made in the first sixteen chapters. The final four chapters of the original, dealing with the practical applications of radar, have been re-written as three chapters which now cover Military Radar Systems, Civil Applications of Radar and Applications of Radar to Physical Science.

Industrial Electronic Engineering

By WELLS L. DAVIS and HERMAN R. WEED. Pp. 514 + x. Sir Isaac Pitman & Sons Ltd., Parker Street, Kingsway, London, W.C.2. Price 55s.

This book is of American origin and is intended for readers having a knowledge of "a.c. theory, the elementary theory of electronic circuits, and elementary differential equations". Roughly half of the book is devoted to gas-filled valves, their characteristics and circuits, rectifiers, both hard-valve and semi-conductor, and photo-cells. In the second half, industrial timing circuits, servomechanisms, voltage regulation, motor control, r.f. heating and X-rays are treated.

Introduction to Ultra-high-frequency Radio Engineering

By STEPHEN A. KNIGHT, F.R.S.A. Pp. 256 + viii. Sir Isaac Pitman & Sons Ltd., Parker Street, Kingsway, London, W.C.2. Price 21s.

The author intends this book as "a guide to new and unorthodox techniques to those who are newly introduced to ultra-high-frequency apparatus" and says that "the reader is assumed to have a reasonable knowledge of 'ordinary' radio techniques and the elementary mathematical treatments which go with them". The ten chapters cover, Introduction, Transmission Lines, Waveguides, Cavity Resonators, Cathode-ray Tubes and Sweep Circuits, U.H.F. Oscillators: Negative and Positive Grid Triodes, Magnetrons and Klystrons, U.H.F. Measurements, Wave Propagation and U.H.F. Aerial Systems.

The treatment is mainly descriptive. The mathematics employed is all fairly simple and it is not used extensively.

The Amplification and Distribution of Sound. (3rd Edition)

By A. E. GREENLEES. Pp. 300 + x. Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 35s.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for August 1954

Date 1954 August	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	− 0·8	+ 2	NM
2	− 0·8	+ 1	NM
3	− 0·8	0	+ 20·5
4	− 0·8	+ 1	NM
5	− 0·8	+ 1	+ 18·1
6	− 0·8	0	+ 16·9
7	− 0·8	+ 2	NM
8	NM	NM	NM
9	− 0·8	+ 2	+ 12·3
10	− 0·7	+ 3	+ 10·8
11	− 0·7	+ 1	+ 8·1
12	− 0·7	+ 2	+ 7·5
13	− 0·7	+ 2	+ 5·6
14	− 0·7	+ 2	NM
15	− 0·7	+ 2	NM
16	− 0·7	+ 3	+ 0·1
17	− 0·7	+ 3	− 1·8
18	− 0·6	+ 3	− 4·2
19	− 0·6	+ 2	− 7·0
20	− 0·6	+ 2	− 9·6
21	− 0·6	+ 1	NM
22	− 0·6	+ 1	NM
23	− 0·6	+ 3	− 16·8
24	− 0·6	+ 1	− 19·3
25	− 0·6	+ 3	− 21·0
26	− 0·6	+ 2	− 23·3
27	− 0·6	+ 3	− 24·5
28	− 0·6	+ 3	NM
29	− 0·6	+ 3	NM
30	− 0·6	+ 3	− 29·7
31	− 0·5	+ 2	− 31·0

The values are based on astronomical data available on 1st September, 1954.

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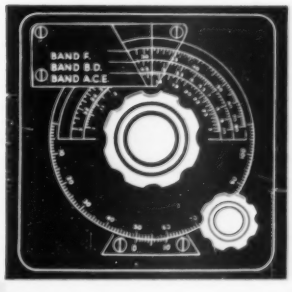

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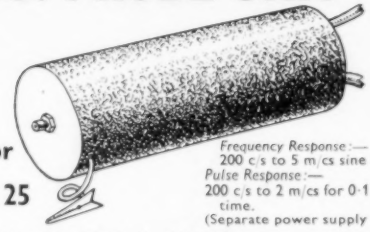

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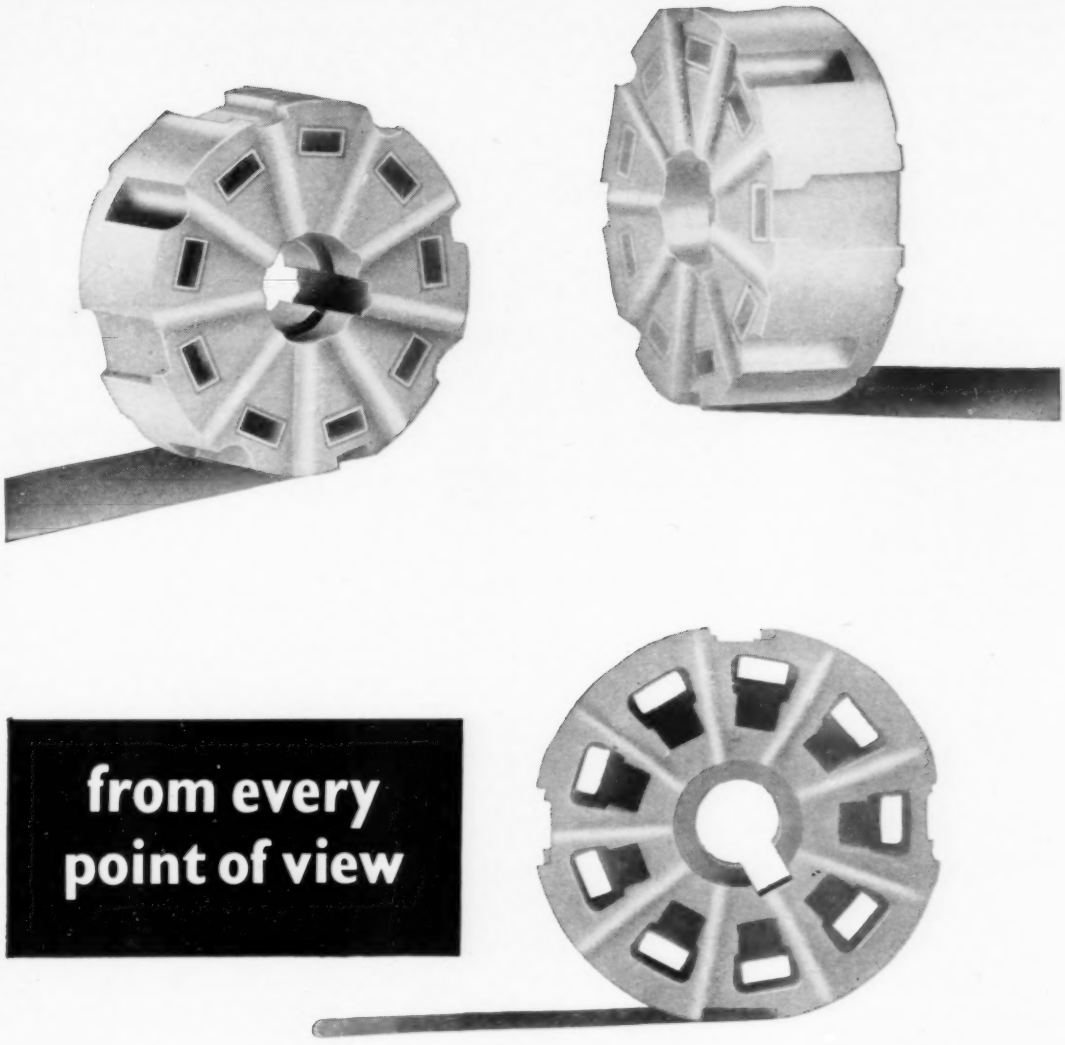

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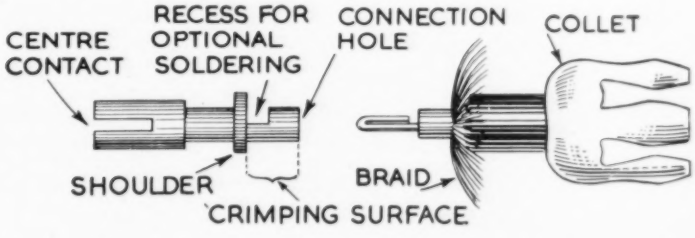

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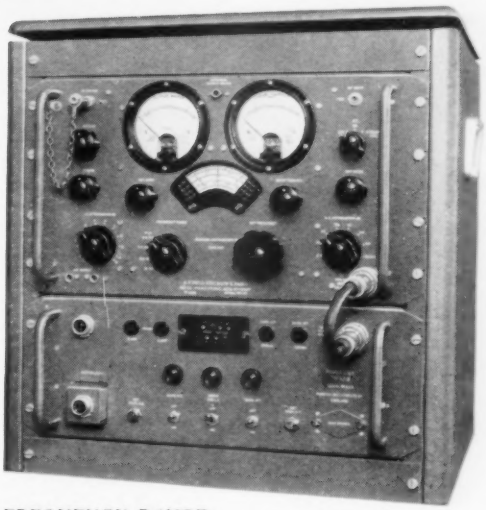

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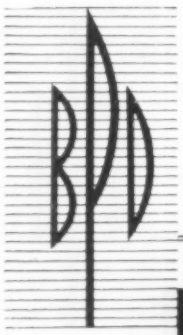

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